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Elektor Electronics

Guitar tuner

Symmetrical power supply

μP-controlled telephone exchange

Medium power AF amplifier

Selective TV preamplifier

Measurement techniques - Part 1

Elektor Electronics

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Front cover
Blind people will soon be able to' read' or listen to instant digital versions of daily newspapers if a new system unveiled at the Royal Institute for the Blind (RNIB) in London is successful.

The Institute's technology department is testing a system that gives blind people access to daily news within hours of publication. At present they must wait for weekly extracts on cassette tape or ask relatives or friends to read the newspaper to them.

In a trial project with The Guardian, text is transmitted over the television network and received in the homes of blind people with access through an authorized screen decoder card in a personal computer. This allows them to 'read' the latest news with the use of a speech synthesizer, as shown, or a transient braille display. The latter option is particularly useful to deaf-blind people whose access to any kind of information is very restricted.
RNIB, 224 Gt Portland St, London W1N 6AA

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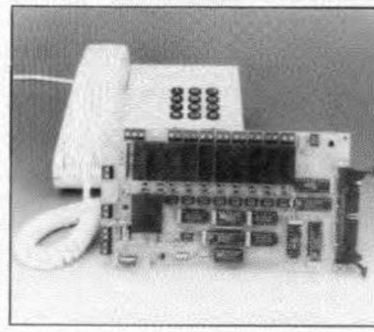

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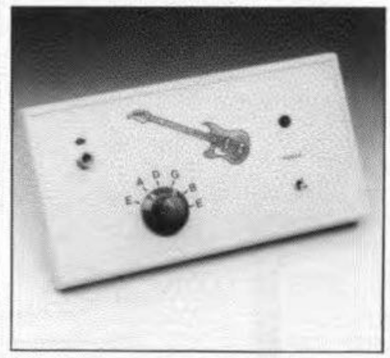

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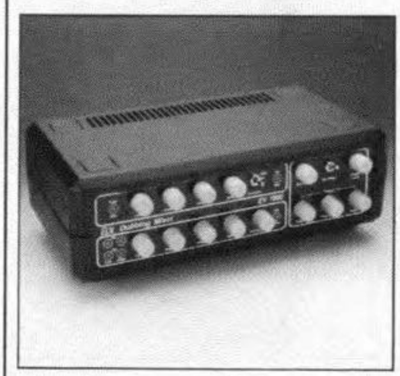

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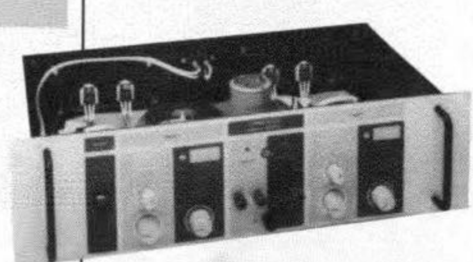

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NEGATIVE RESISTANCE

by Dr. Ir. A.H. Boerdijk

A portion of the current-voltage characteristic of certain devices, such as the thyristor, the tunnel diode and the magnetron, has a negative slope, that is, the current decreases with increasing applied voltage or the voltage drops when the current increases. This behaviour is, of course, opposite to that of an ohmic resistance. Whereas an ohmic (positive) resistance consumes power, a negative resistance appears to supply power. Negative resistance may be simulated electronically as described in this article.

From a pure arithmetic point of view, negative resistance remains resistance with the only difference that it is preceded by a minus sign. Figure 1 shows a conventional ohmic resistance and a negative resistance with an identical voltage applied across them. The difference in behaviour of the two is clear: the currents through them flow in opposite directions.

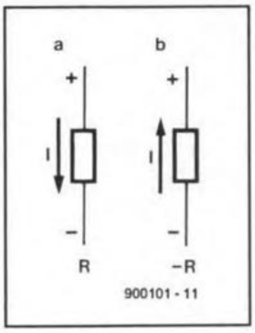

Fig. 1.

When a positive and a negative resistance are connected in series or parallel as shown in Fig. 2, the results are very interesting. The

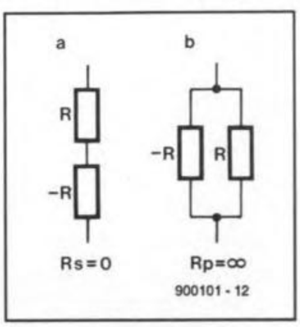

Fig. 2.

series combination (Fig. 2a) yields a short circuit:

$R_s = R + (-R) = 0.$

The parallel network (Fig. 2b) yields

$R_p = -R^2 / [R + (-R)] = -R^2/0 = -\infty,$

that is, a perfect insulator.

To confuse you further, in Fig. 3 a 10-V potential is connected across a series combination of a positive resistance of 1001 kΩ and a negative resistance of 1000 kΩ.

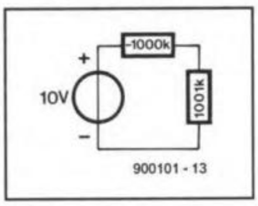

Fig. 3.

The total resistance in the loop, ignoring the internal resistance of the voltage source, is 1 kΩ. The current flowing in the loop is therefore 10 mA, and this causes a drop of 10.01 kV (!) across the 1001 kΩ resistance.

This does not indicate a new way of generating very high voltages, of course, as a quick consideration of the power distribution shows.

In Fig. 1a, the positive resistance **dissipates** a power $P = I^2 R$ or $P = U^2/R$, whereas in Fig. 1b the negative resistance **delivers** power to the voltage source. This means that negative resistance is not just a passive component and also that it can not exist by itself (since the power delivered to the voltage source must come from somewhere).

In fact, a negative resistance may be simulated by an electronic network as shown in

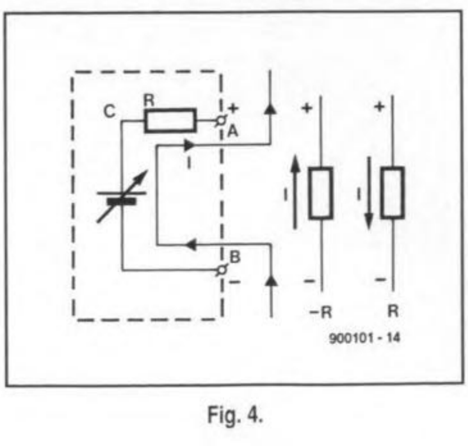

Fig. 4.

Fig. 4, where it exists between A and B. Terminal A is connected to a variable voltage source between terminals C and B that generates a voltage $U_{CB} = 2U_{AB}$. If the potential at A is positive with respect to B, the voltage at C is so, too. A current, $I = U_{CA}/R$ flows through R in the direction indicated, that is, from B (–) to A (+). In other words, the resistance between terminals A and B is negative.

When considering the operation of this network, it is important to pay attention only to terminals A and B: the circuitry hidden behind them is of no consequence here.

In practice

The circuit in Fig. 5 is constructed from an opamp and three resistors, while a negative resistance of –1 kΩ is simulated between terminals A and B. The operation may be checked by connecting a 4.7 kΩ resistor in series with terminals A and B. The total re-

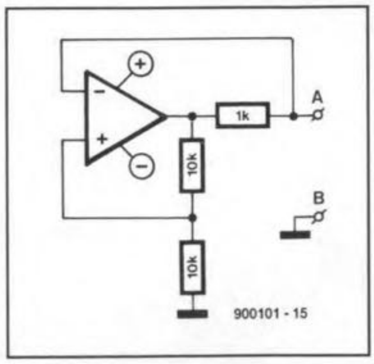

Fig. 5.

sistance measured with a standard ohmmeter is 3.7 kΩ, which shows that the effect of a negative resistance can be measured. The value of it depends on the value of the output resistor used in the simulation circuit and the ratio of the other two resistors. Replacing the fixed 1 kΩ resistor by a variable type enables a wide range of negative resistance values to be obtained.

Another fairly simple method is to connect a conventional resistor in series with the negative resistance. In this example, this resistor should have a value not exceeding 1 kΩ to prevent the negative resistance from disappearing.

If a resistor (fixed or variable) greater than 1 kΩ is connected in parallel with terminals A and B, the negative resistance increases (becomes more negative). The circuit in Fig. 5 is very suitable for experimenting with negative resistance.

When the circuit of Fig. 5 is translated into a practical design, a certain load, R_v, will exist between terminals A and B. This load has an effect on the operation of the circuit and its value must therefore be higher than the absolute value of $-R$, that is, in this circuit greater than 1 kΩ.

If the load across terminals A and B is always smaller than $-R$, the circuit is still usable, but the connections to the inputs of the opamp must be reversed (this maintains the required feedback).

Although the circuit in principle becomes unstable only when the numerical values of R_v and $-R$ are identical, it will be found in practice that it does not function satisfactorily when the values are close to another.

It will have become clear that the maximum potential drop across the negative resistance is highly dependent on the voltage source used for the simulation circuit. This also explains why the circuit of Fig. 3 does not generate a very high voltage, although it works satisfactorily: the supply voltage is not high enough.

The output characteristic of the opamp determines the maximum current that can flow through the negative resistance. If larger currents are wanted, the output of the opamp must be provided with an additional stage. It is, of course, also possible to use an opamp that handles larger currents.

Applications

In practical electronics, negative resistance is used to compensate (ohmic) losses. A typical example is an *LC* circuit as shown in Fig. 6. The resonant frequency of this is 800 Hz and the *Q*-factor is 5.4. The value of *Q* is low, because it is heavily affected by the (loss) resistance of the inductor. It may be improved considerably by adding a variable negative resistance in parallel with the circuit. This is accomplished as shown: the fixed negative resistance is connected between A and B, and the potentiometer enables the losses caused by the resistance of the inductor to be compensated.

It is even possible to set the circuit into oscillation by making the negative resistance sufficiently large, that is, by reducing the value of the parallel resistance. The frequency range of the circuit will then be restricted, however, by the bandwidth of the simulation circuit.

Another application is the improvement of the control range of small d.c. motors. The rotating speed of such motors, especially at the low end of the range, is heavily dependent on the load moment. In fact, at a given point the motor just stops abruptly.

This behaviour may be improved greatly with the aid of the circuit in Fig. 7, which contains not only a variable negative resistance but also a variable supply for the motor. Potentiometer P1 controls the rotating speed of the motor, while P2 sets the value of the negative resistance.

Experiments with a small d.c. motor showed that the deviation of the moment vs speed characteristic from the ideal could be improved by a factor of 2.7.

A final application is the use of a 3-Ω negative resistance to charge a battery. Connected to a 12-V battery, the charging current is 4 A; connected to a 6-V battery, the charging current is 2 A.

Such a negative-resistance charger has some peculiar properties: the connections to the battery terminals may be reversed with impunity and the short-circuit current amounts to nought. ■

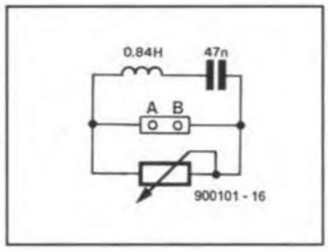

Fig. 6.

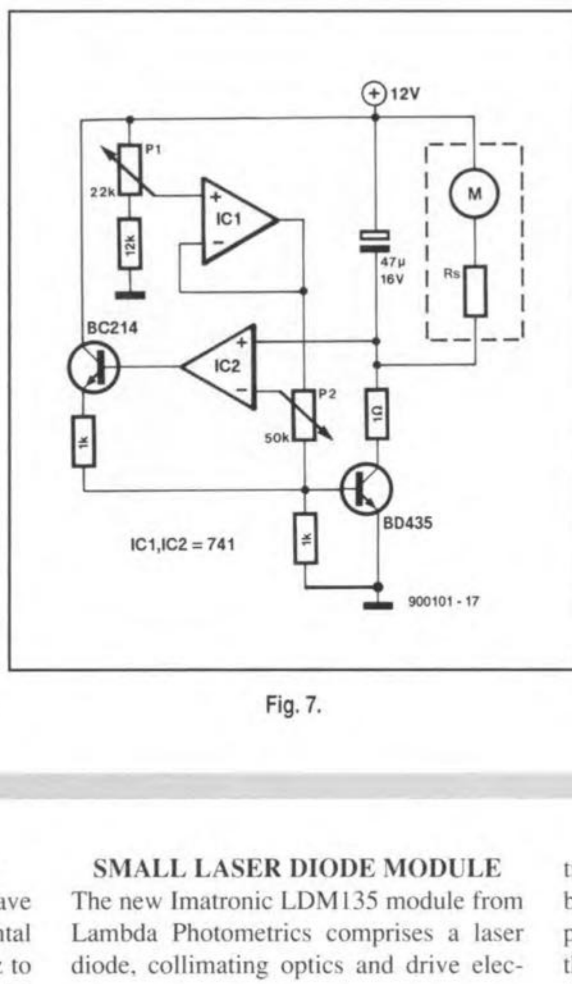

Fig. 7.

LOW-FREQUENCY CRYSTALS

Euroquartz of Crewkerne, Somerset, have available low-frequency, fundamental mode crystals in the range 40–800 kHz to customer-specified frequencies. Crystal cuts include +5° X-cut, DT-cut and CT-cut. Holder types include HC49/U, HC50/U, HC51/U, HC6/U, HC13/U andHC34/U, all in resistance-weld packaging.

Calibration tolerances are available from ±20 ppm and with temperature stability characteristics dependent on frequency and operating temperature range.

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SMALL LASER DIODE MODULE

The new Imatronic LDM135 module from Lambda Photometrics comprises a laser diode, collimating optics and drive electronics in a compact housing of 16 mm dia. by 49 mm height. Power (3–5.25 V) is applied via two colour-coded flying leads. Although preset to produce a parallel beam, the focal length is easily adjusted to focus the beam to a fine spot.

There are four basic versions: 0.5 mW, 1 mW, 2 mW and 3 mW at 660–685 nm.

The unit makes an ideal replacement for He:Ne lasers and in many applications will prove more suitable.

Unit prices start at £195, but quantity discounts are available.

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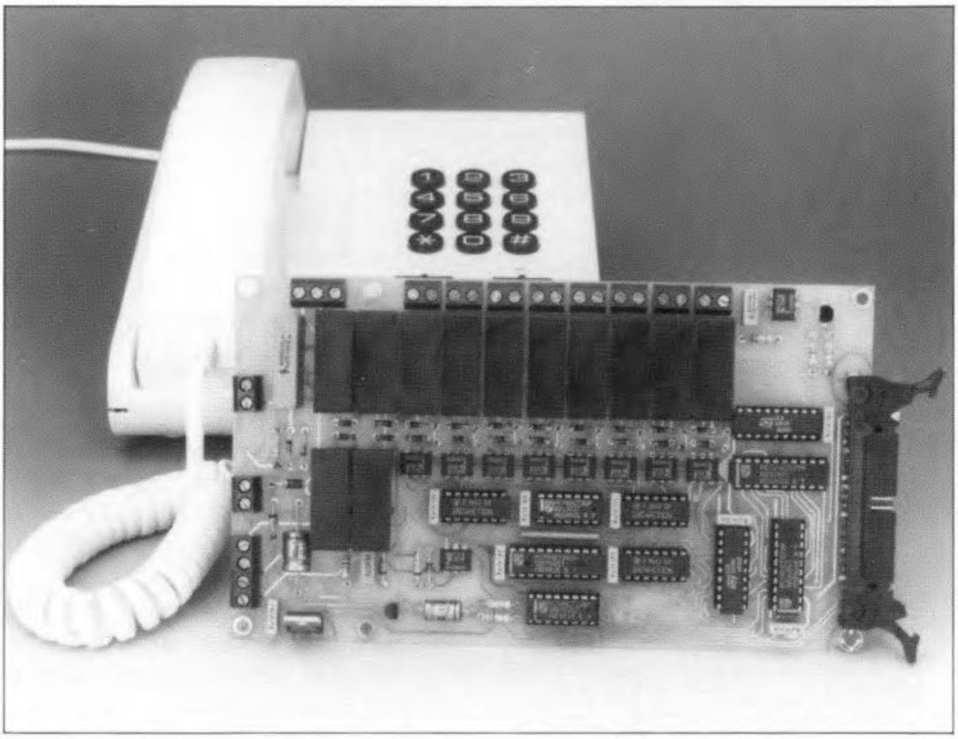
MICROPROCESSOR-CONTROLLED TELEPHONE EXCHANGE

The telephone exchange presented here allows up to eight pulse-dialling telephone sets to be connected, and has an option for connecting calls to or from an external (trunk) telephone line. The unit is controlled by the popular 8052-based BASIC computer we introduced a few years ago.

A. Rigby



MAIN FEATURES

- 8 internal lines
- 1 external line
- memory for 10 numbers
- internal through connections
- versatile computer control
- automatic hold for external line
- simple-to-extend
- can be interfaced to a PC
- selective external call acceptance
- shortcut dial codes for external number
- works with pulse-dialling telephone sets
- one optional relay for extra switching function

Since the telephone exchange is controlled by a computer, it is relatively easy to add or change certain features simply by extending or changing the control program. The nice thing about the 8052-based computer used here (Refs. 1 and 2) is that it can be programmed in BASIC, a computer language familiar to many. In the present application, the BASIC computer runs its application program from an on-board EPROM. All that is required to modify this program is a terminal or a PC running a communications program, and a three-wire link to the BASIC computer. With these tools, the user is at liberty to edit and extend the existing control program in order to 'customize' the telephone exchange. The terminal or PC is no longer required once the control program has been tested and found to work all right. If you have no intention to change the 'standard' control program, or lack the ability to program in BASIC, simply use the ready-programmed EPROM available for this project. In most cases, this standard control program will provide all the necessary functions users of a telephone exchange for a small network in the home or small office or workshop have come to expect.

Telephone: the basics

Before discussing the operation of the telephone exchange, it is useful to look at the basic operation of the telephone system. In the following discussion, it is assumed that pulse-dialling telephone sets are used. The operation of tone-dialling (DTMF) is not covered. Details on this system may be found in Ref. 3.

Figure 1 shows the general lay-out of a telephone connection. When the receiver is on the hook, the bell inside the telephone set is connected to the telephone line. When the receiver is lifted, the voice circuit of the set is connected to the telephone network, and a direct current flows through the microphone. The telephone extensions connected to the network receive their supply voltage from the local telephone exchange. All sets are connected to two lines and operate free from the earth line. The use of balanced lines is a simple, yet effective, way, to eliminate noise in the network. Since any noise induced on the network is, in principle, equally strong and of equal phase on the 'a' and 'b' lines, it is effectively inaudible.

Outgoing calls

The timing diagrams in Fig. 2 show the switching sequences during a telephone call. Again, only the 'a' and 'b' lines are involved in establishing the call. Normally, a voltage of 50 to 60 V exists between these lines. The exchange detects that a receiver is lifted when the line voltage drops to about 10 V, and a microphone current of about 20 mA is established. Next, the exchange sends the dial tone to the calling extension to indicate that a number may be dialled. In the pulse-dialling system, the current loop is interrupted repetitively. The pulse rate usually lies between 9 and 11 pulses per second. The 'break' period is called 'pulse', and the 'connect' period is called 'pause'. The pulse length is generally defined as 61.5% ±3% of the period. Assuming that the period is 100-ms, the current is interrupted for periods of 58.5 to 64.5 ms. The pause allowed between successive numbers is 0.7 to 1 s.

The local exchange starts to call up the wanted extension with the aid of a ringing signal after the complete number has been received from the calling extension. When the call is answered, the exchange starts to put a cost count signal on the 'a' and 'b' lines. This signal is a sine-wave burst with an amplitude of about 50 V. Since it is the same on the 'a' and 'b' line, it is inaudible to the calling as well as to the called party. A cost counter, however, is connected asymmetrically to the network to allow it to detect the pulses. When either party rings off (puts the receiver down), the voltage between the 'a' and 'b' line reverts to the 'standby' level of 50 to 60 V.

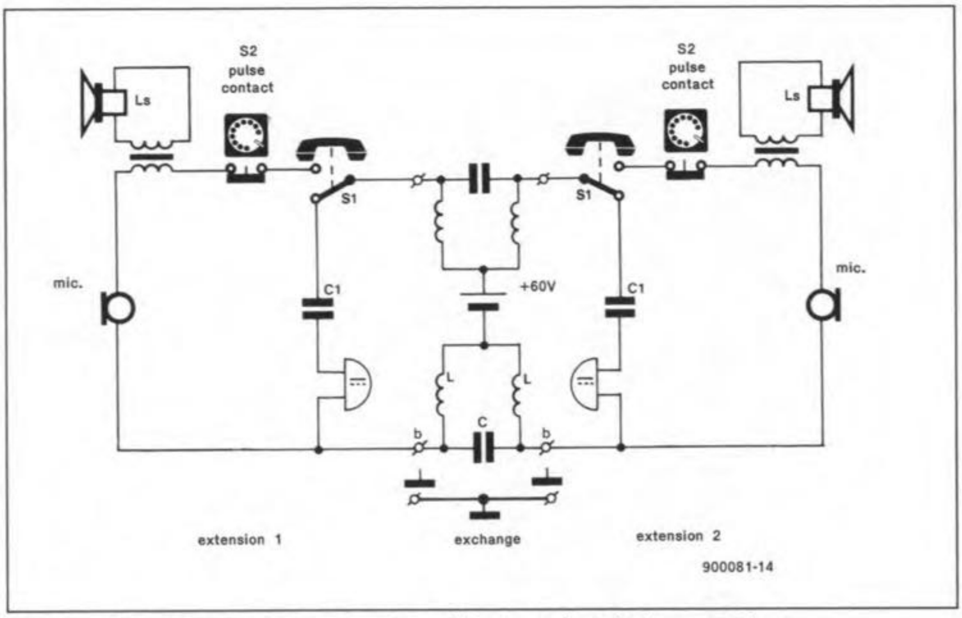


Fig. 1. Illustrating the basic operation of the two-wire telephone system.

Incoming calls

The operation of the telephone system in the case of incoming calls is illustrated in Fig. 3. An incoming call is detected by the ringing signal produced by the telephone set. The exchange calls up the extension by putting an alternating voltage of about 50 V_{PP} on the 'a' and 'b' lines. The fact that the signals on 'a' and 'b' are in anti-phase allows the telephone to detect the ring signal and actuate a sounder device (usually a small bell or buzzer). The ringing continues until the called party lifts the receiver to answer the call. If the call is not answered after a predetermined number of rings, the connection is broken (in the exchange discussed here, the maximum number of rings is set to 13). When the called party lifts the receiver before the last ring, the previously mentioned direct current flow is established, enabling the exchange to detect that the call is answered. The telephone conversation can begin!

Electronics at work

The signal sequences shown in Figs. 2a and 2b are generated and processed by the interface board of the telephone exchange, while the control functions are carried out by the BASIC computer. The function of the interface board, of which the circuit diagram is shown in Fig. 3, is to convert the digital signals supplied by the computer board to telephone network signals, and vice versa.

The eight interfaces that establish the connections with the telephone extensions are shown at the top of the circuit diagram. The extensions are connected either to the WAIT line or to the VOICE line. Extensions used for a telephone conversation are always connected to the VOICE line, which provides the necessary supply voltage. Extensions not involved in the conversation are connected to the WAIT line, and produce the 'engaged' tone when the receiver is lifted. A number can only be dialled when the exchange is back in the wait state with all extensions connected to the VOICE line.

The interface board is linked to the BASIC computer via connector K14, which carries all the necessary signals for proper communication between these units. The $\overline{Y7}$ signal supplied by the address decoder on the BASIC computer board is used to select the logic on the interface board. The line is actuated in the address range between $E000_H$ and $FFFF_H$, which is split into three parts with the aid of address lines A10, A11 and A12, giving buffer devices IC15, IC16 and IC17 their

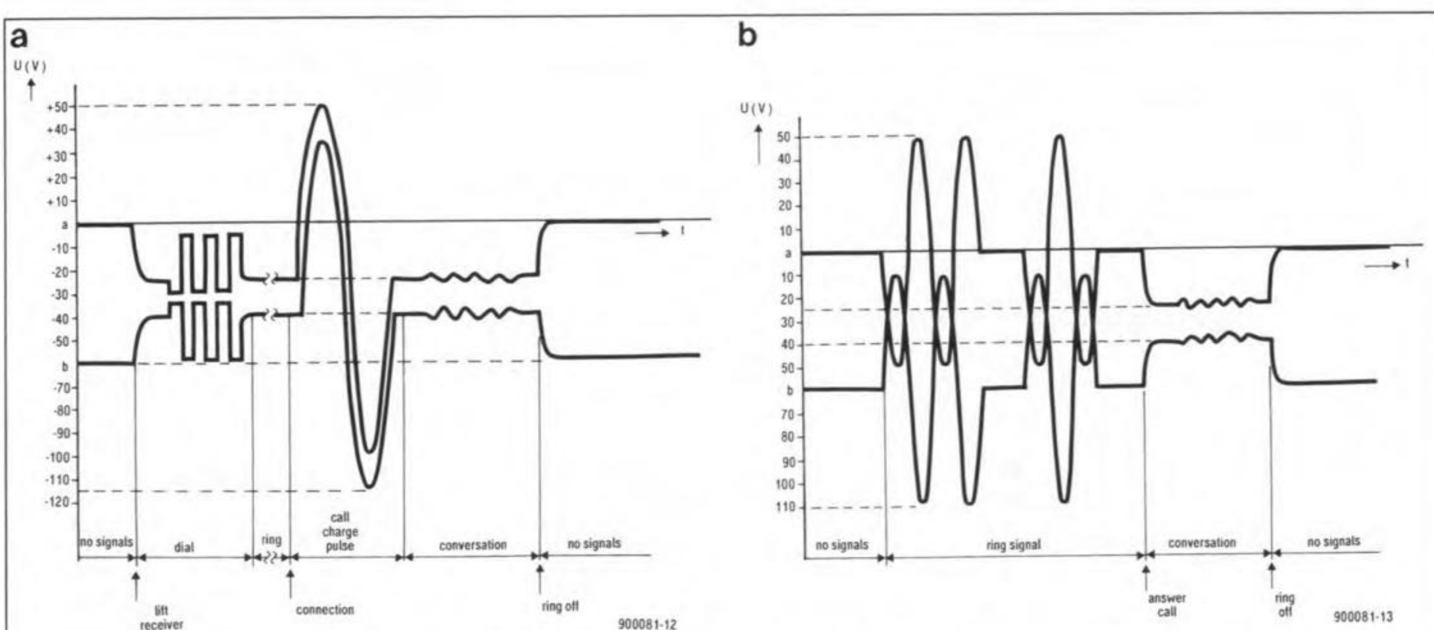


Fig. 2. Waveform sequences on the telephone lines, showing the call charge pulses (Fig. 2a) and the ring signal (Fig. 2b).

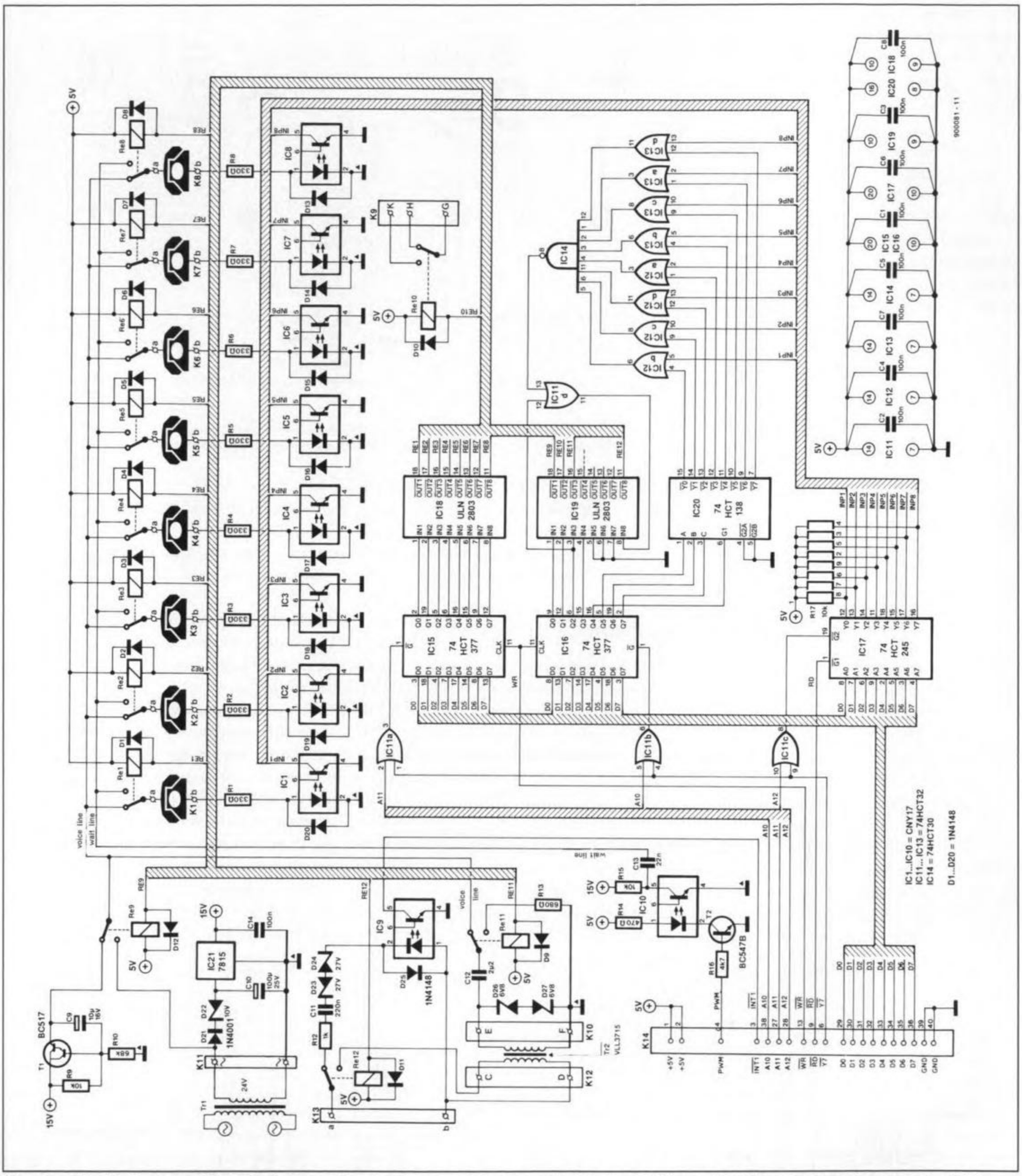


Fig. 3. Circuit diagram of the telephone exchange. This circuit is connected to the 8052-based BASIC computer via connector K14.

proper location in the memory map. The $\overline{\text{INT}}$ signal supplied by the interface board serves to wake up the BASIC processor from its stand-by state when a ringing signal is detected on the external line. The 8052 processor generates the 'engaged' tone on the PWM line. A dial tone is not generated—the network is free for dialling an extension when the receiver is silent upon being lifted. The remaining lines on K14 carry data signals, read and write signals, and the supply voltage.

Circuits IC15 and IC16 are latches that function as additional I/O registers for the control of the switching functions available in the exchange. Relays are used for the actual switching actions. Eight-bit register IC15 controls relays Re1–Re8 via the power drivers contained in IC18. These relays are used to switch the associated telephone sets between the VOICE and the WAIT line. The three least-significant datalines on IC16 switch relays Re9, Re10 and Re11. The first, Re9, is used to generate the ringing signal. In the stand-by state, transistor T1 is connected to the VOICE line, and provides all telephone sets with their supply voltage via the VOICE line. The gyrator configuration of the transistor prevents the supply short-circuiting voice signals from being superimposed on the direct voltage. When Re9 is switched, the full transformer voltage is applied to the VOICE line. As a result, the bell in the extension connected to the VOICE line starts to ring. The calling party hears the ringing signal as a series of buzzing tones.

Relay Re10 is intended for optional extensions, such as a telephone-controlled door opener, and can be controlled by appropriate modifications to the BASIC control program.

Relay Re11 is used to transfer a call received on the external line to another extension in the network. By switching Re11, the external line is terminated at the required impedance. As a result, the line is held while the exchange is being used for internal calls.

Making a call

When the receiver on any of the extensions is lifted to make a call, a current starts to flow that causes the LED in the associated opto-coupler to light. This results in the relevant $\overline{\text{INP}}$ line being pulled low. The processor identifies the calling extension by reading the logic 0 it produces in IC17 at address $EFFF_H$. Next, a write command is issued to IC15 and IC16 (at addresses $FBFF_H$ and $F7FF_H$ respectively) to connect all other extensions to the WAIT line. These extensions are effectively disabled and produce the 'engaged' tone when the receiver is lifted.

The processor counts the dialling pulses produced by the calling extensions via IC17. The dialled number determines what happens next. When a 0 is dialled, relay Re12 is actuated, and the external line is selected to establish a connection to another telephone network or another exchange. The line transformer, Tr2, is connected to the external line, and all dialling pulses that follow the 0 are fed to the external line by Re12 being actuated in their rhythm. The relay contact switches between a low impedance (the line transformer) and a high impedance (the ring pulse detector). The dialling pulses are fed out of the exchange via IC11, IC12, IC13 and IC14, after the right OR gate (IC12a–IC12d or IC13a–IC13d) has been enabled via IC20. Gate IC11d ensures that dialling pulses produced by one of the internal extensions are not passed to the external line while this is on hold. This is an important feature when an call received via the external line is being transferred to another extension served by the exchange.

Receiving external calls

Calls that reach the exchange via the external line are detected by the ring pulse detector based around D23, D24, D25, R12, C11 and IC9. When a ringing pulse is detected, IC9 pulls the $\overline{\text{INT1}}$ line of the BASIC processor logic

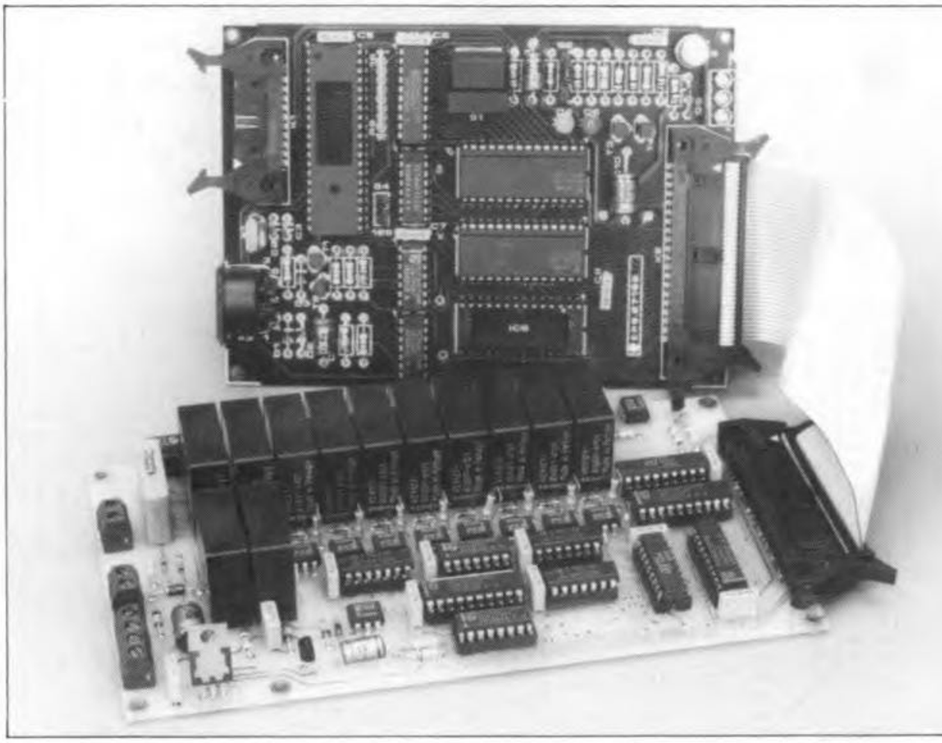


Fig. 4. Ready for use: completed BASIC computer and telephone exhange boards.

low. Only those extensions allowed to accept calls from the external line remain on the VOICE line; all others are connected to the WAIT line. A ringing signal is placed on the VOICE line with the aid of Re9. A total of 13 rings with 2.5-second pauses is allowed. The first extension that answers the call is connected to the external line. Once again the telephone conversation can begin!

After it has been answered, the external call can be transferred to another internal extension. To do this, the active extension puts the receiver down and dials the number of the wanted extension. The external line is not disconnected until any receiver has been on the hook for more than five seconds. While the external caller is on hold, the answering extension dials another extension. The external line is connected to whichever extension remains on the line when the other puts the receiver down. If the wanted extension does not answer the call, another one may be tried. In all cases, however, the total time the receiver is down must not exceed five seconds. If none of the other extensions answers, the external caller may be connected again by dialling your own number.

Construction

Figure 5 shows the track lay-outs and the component mounting plan of the double-sided, through-plated printed circuit board for the telephone exchange. The board has been designed to form a compact unit together with the BASIC computer. The greater part of the board space is reserved for the relays and the opto-couplers. Assuming that the ready-made board is used, the actual construction is unlikely to present problems if carried out with the necessary care. Accurate soldering is a must, though, to prevent short-circuits.

The two transformers are fitted as external parts on separate pieces of veroboard or stripboard to keep the overall size (and with it the cost) of the interface board as small as possible. Be sure to observe the necessary safety precautions because of the presence of the mains voltage on the mains transformer board. The BASIC computer is powered by a separate, regulated, 5-V supply.

The telephone sets and the transformers are connected to plastic or ceramic terminal blocks fitted on the PCB. The contacts of (optional) relay Re10 are available on connector K9 for experimental purposes.

The construction and operation of the BASIC computer is not covered here—for details, please consult Refs. 1 and 2. A small modification must be made to the existing circuit in regard of signals PWM, $\overline{\text{Y7}}$ and $\overline{\text{INT1}}$, which are not available on the expansion connector of the computer board. Three wire links are fitted to overcome this problem: connect pin 3 of K2 ($\overline{\text{INT1}}$ signal) to pin 10 of K1. Next, connect pin 4 of K2 (PWM signal) to pin 15 of K1. Finally, connect pin 8 of K2 ($\overline{\text{Y7}}$ signal) to pin 7 of IC3 (74HCT138). Since these wires go to previously unused pins of K2, they do not affect the normal operation of the BASIC computer.

The two boards are connected via a short length of flatcable fitted with IDC sockets that connect to K2 at the BASIC computer side and K14 at the interface board side. After

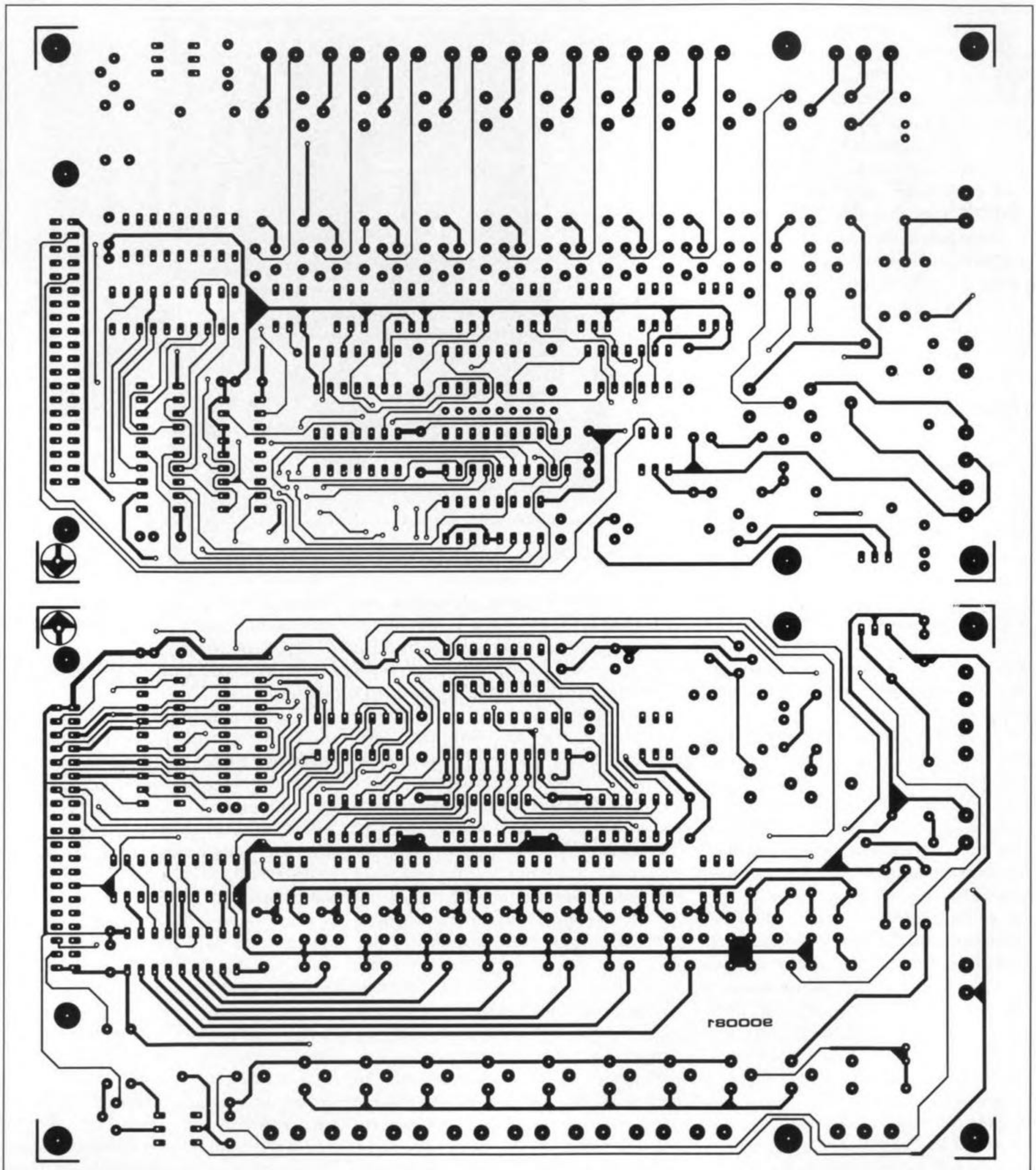


Fig. 5a. Track layouts of the double-sided, through-plated printed-circuit board.

fitting the system EPROM into its socket on the BASIC computer board, and resetting the system, the exchange is ready for use.

Control software

Software is essential for any microprocessor-based system. The control program for the telephone exchange is written in BASIC with plenty of comment in the listing to explain the operation. As already stated, the program is supplied in the form of an EPROM. Those of you who want to change it may get out their terminal or PC, connect it to the BASIC computer, and suspend the program by typing control-C. Next, LIST the program, and edit it as required. RUN the program to check that it does what you want. The syntax requirements of the 8052 BASIC interpreter are covered in the relevant Intel manual, while possible problems with the communication between the terminal or PC and the BASIC computer are tackled in Refs. 1 and 2.

A short description is given of the function of the main routines in the control program:

- the internal numbers start with a '1', i.e., the extensions in the network have numbers 11 up and including 18.

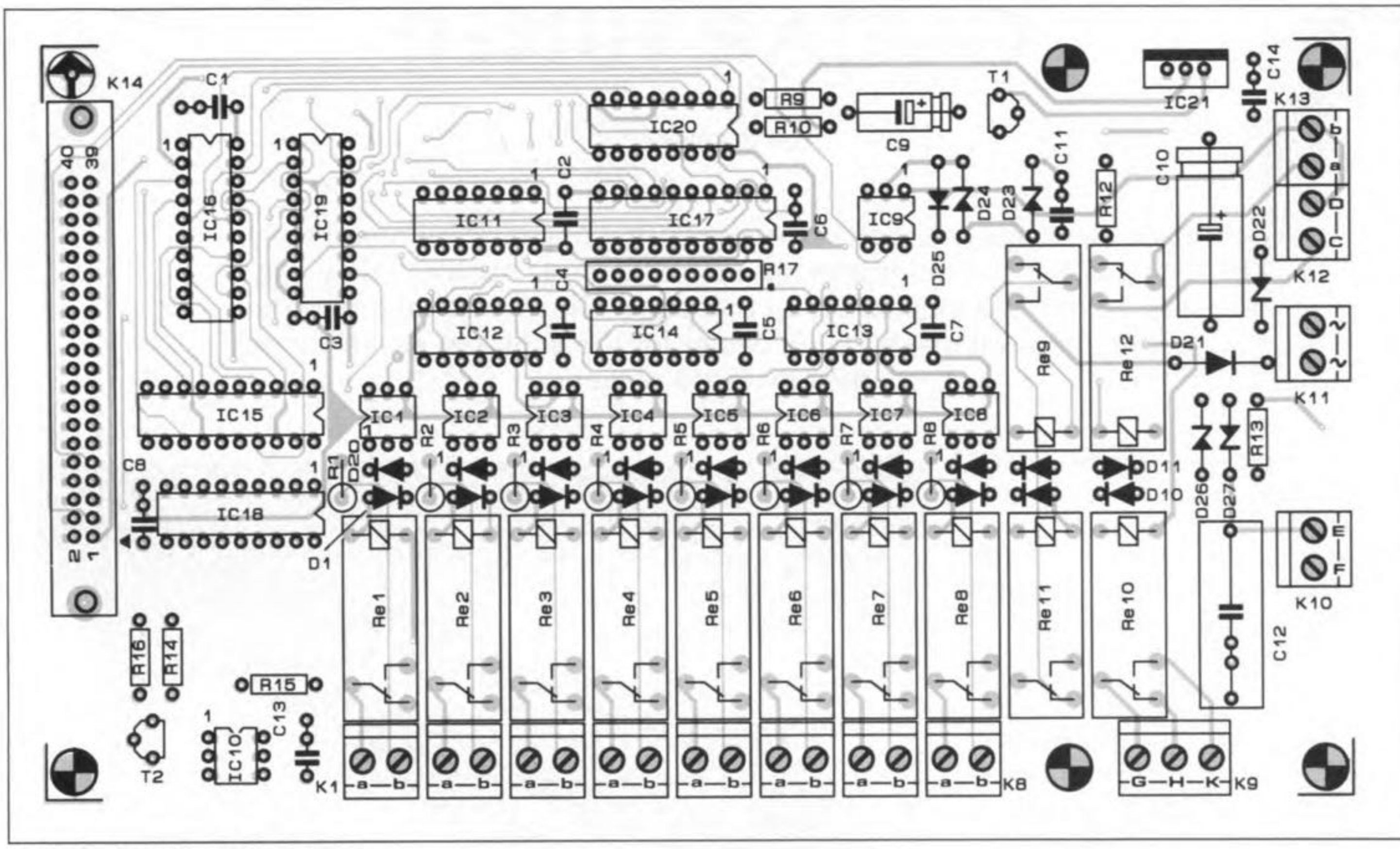


Fig. 5b. Component mounting plan.

COMPONENTS LIST

Resistors:

8	330Ω	R1–R8
2	10kΩ	R9;R15
1	68kΩ	R10
1	1kΩ	R12
1	680Ω	R13
1	470Ω	R14
1	4k7	R16
1	SIL array 8×10kΩ	R17

Capacitors:

9	100nF	C1–C8;C14
1	10µF 16V	C9
1	100µF 25V	C10
1	220nF	C11
1	2µ2 100V	C12
1	22nF	C13

Semiconductors:

10	CNY17	IC1–IC10
3	74HCT32	IC11;IC12;IC13
1	74HCT30	IC14
2	74HCT377	IC15;IC16
1	74HCT245	IC17
2	ULN2803	IC18;IC19
1	74HCT138	IC20
1	7815	IC21
1	BC517	T1
1	BC547B	T2
21	1N4148	D1–D20;D25
1	1N4001	D21
1	10V 0.4W zener diode	D22
2	27V 0.4W zener diode	D23;D24
2	6V8 0.4W zener diode	D26;D27
1	EPROM with control program for BASIC computer. Order code ESS 5941.	

Miscellaneous:

12	2-way PCB terminal block	K1–K8; K10–K13
1	3-way PCB terminal block	K9
1	40-way PCB-mount plug with eject headers	K14
12	5-V SPDT PCB-mount relay, Siemens type V23127-B0001-A101	Re1–Re12
1	24V 3.3VA mains transformer	Tr1
1	telephone line transformer type VLL3715T	Tr2
1	printed-circuit board	900081

- a total of ten shortcut codes is allowed for external numbers. These codes start with a '2', i.e., 20 up to and including 29 are available.
- the codes used for shortcut dialling are stored by first dialling '3'. Next, dial the code (0–9), followed by the number of the external connection. The processor stores the shortcut code and the associated number when the receiver is put down. All codes are available to all extensions in the network served by the exchange, and they may be changed at any time by any extension.
- a particular extension can be disabled from receiving external calls by dialling '5'. This can be undone by dialling '6'.

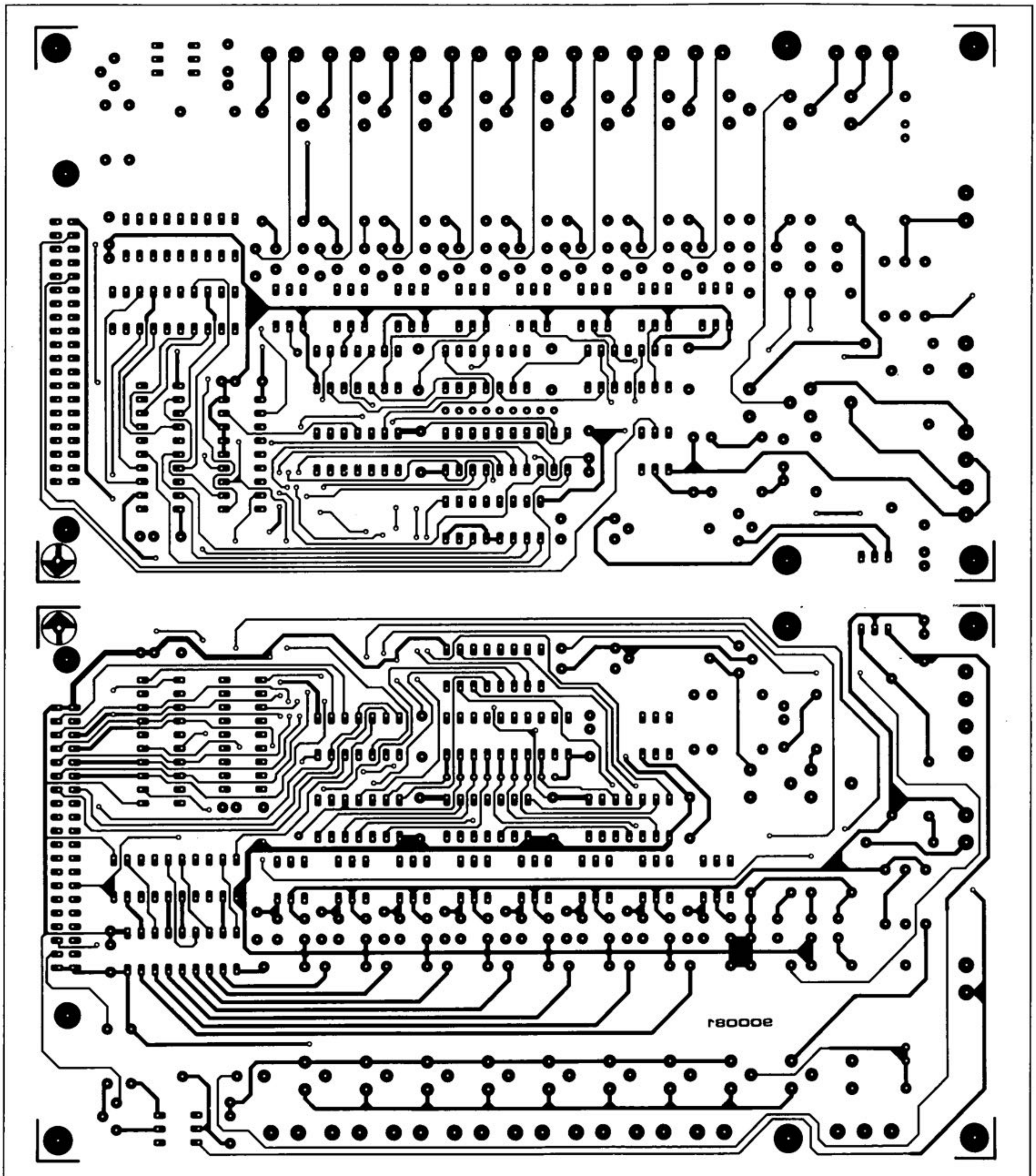
The function of dial numbers '4', '7', '8' and '9' is not fixed, although the software has built-in routines to intercept them. Number '8', for instance, could be used to switch on relay Re10, and number '9', to switch it off again. To be able to do this, you have to include the appropriate write command in the number interception subroutine, test the option, and program a new EPROM. ■

References:

1. BASIC computer. *Elektor Electronics* November 1987.
2. ROM-copy for 8052-BASIC computer. *Elektor Electronics* September 1990.
3. Dual-tone multi-frequency (DTMF) decoder. *Elektor Electronics* May 1989.

IMPORTANT NOTICE

The telephone exchange described here is not type-approved by British Telecom and may not be connected to the public switched telephone network (PSTN). In countries other than the UK, the relevant PTT authorities should be contacted about type-approval. In the text, 'external line' is meant to indicate a wire system other than the PSTN.


900081

CTIONS CORRECTIONS CORRE

400-watt laboratory power supply

October 1989 and November 1990

A number of constructors of this popular project have brought the following problems to our attention.

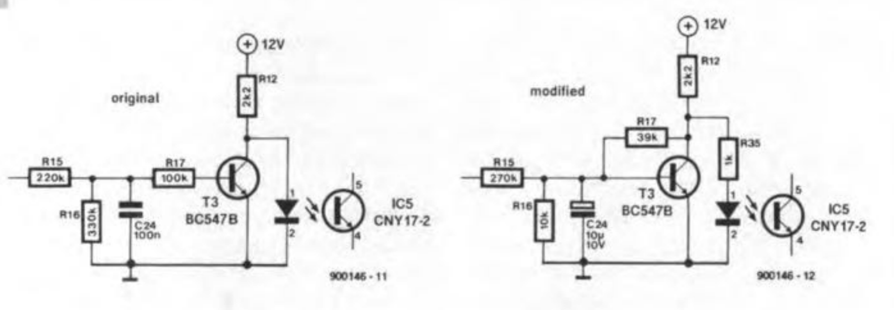
1. The onset point of the current limit circuit lies at about 3 A, which is too low. Solve this problem by replacing T1 with a Type BC517 darlington transistor, and R20 with a 82kΩ resistor.

2. Depending on the current transfer ratio of the optocoupler used, the transformer produces ticking noises. This effect, which is caused by overshoot in the pre-regulation circuit, may be traced with the aid of an oscilloscope monitoring the voltage across C26 at a moderate load current. The capacitor must be charged at each cycle of the mains frequency, and not once every five cycles. The problem is best solved by reducing the amplification of the regulation circuit. Replace R17 with a 39 kΩ resistor, and create feedback by fitting it between the base and the collector of T3. Also add a resistor in series with the optocoupler. These two changes are illustrated in Figs. 1 and 2. Lower R16 to 10 kΩ, increase C24 to 10 µF, and increase R15 to 270 kΩ.

3. Excessive heating of the transformer is caused by a d.c. component in the primary winding. This is simple to remedy by fitting a capacitor of any value between 47 nF and 470 nF, and a voltage rating of 630 V, across the primary connections. This capacitor is conveniently mounted on to the PCB terminal block that connects the transformer to the mains.

4. One final point: when using LED DVMs for the voltage/current indication, their ground line must be connected to the **positive** terminal of C12.

1



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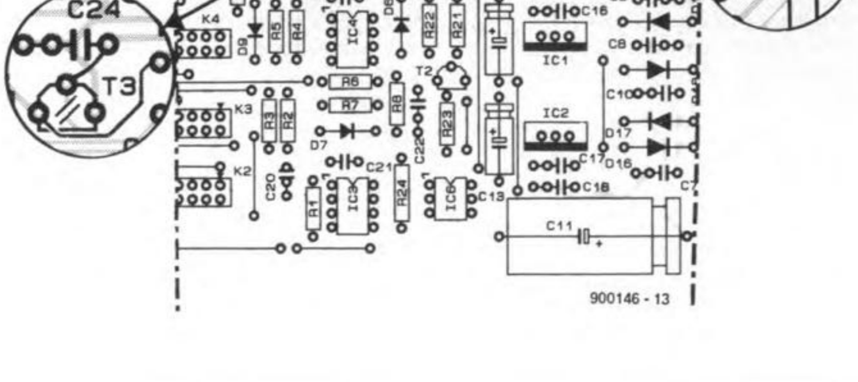

Hard disk monitor

December 1989

In some cases, the circuit will not reset properly because the CLEAR input of IC3A is erroneously connected to ground. Cut the ground track to pin 3 of IC3, and use a short wire to connect pin 3 to pin 16 (+5 V).

Microprocessor-controlled telephone exchange

October 1990

In some cases, the timing of the signals applied to IC17 causes a latch-up in the circuit, so that the exchange does not detect the state of the connected telephones properly. Solve this problem by cutting the track to pin 1 of IC17, and connecting pin 1 to ground (a suitable point is the lower terminal of C6).

The text on the fitting of wires on the BASIC computer board (page 19, towards the bottom of the right-hand column) should be modified to read: 'Finally, connect pin 6 of K2 to pin 7 of IC3 ($\overline{Y7}$ signal).'

S-VHS/CVBS-to-RGB converter (2)

October 1990

The capacitor marked 'C37', next to R21 on the component overlay (Fig. 7b and ready-made printed circuit board), should be marked 'C39'.

In case they are difficult to obtain locally, inductors type 119-LN-A3753 (L1) and 119-LN-A5783 (L2) may be replaced with the respective types 119-ANA-5874HM and 119-ANA-5871HM, also from Toko, Inc. Suggested suppliers are Cirkit Distribution Ltd., and C-I Electronics.

EPROM simulator

December 1989

Counters IC3 and IC4 may not function properly owing to a too low supply voltage. This problem may be solved by replacing IC12 with a 7806. Alternatively, use BAT85 diodes in positions D1 and D2.

Programmer for the 8751

November 1990

The ready-programmed 8751 for this project is available at £35.25 (plus VAT) under order number **ESS 7061**, not under order number ESS 5951 as stated on the Readers Services pages in the November and December 1990 issues.

GUITAR TUNER

by T. Giffard

A guitar is one of those instruments that needs to be tuned daily. The number of tuning aids on the market indicate that many players do not find this tuning all that easy. The electronic tuner presented here will make tuning easier for the beginner as well as for the advanced player.

For almost three centuries guitarists have used a tuning-fork (invented in 1711 by John Shore) or a pitch-pipe (which appeared later that century) as an aid to tuning their instrument. Between the two world wars, electrical aids to tuning became popular. Typical of these is the resonoscope, introduced almost simultaneously in the USA and the UK in 1936, which gives a visual indication at the correct tuning pitch. Some drawbacks of them are that they are not easily made and that beginners never learn how to tune the guitar properly. The tuning-fork and pitch-pipe (also known as flue) are not ideal either. The tone of a tuning-fork is barely audible and lasts but for a few moments. The pitch-pipe has a better volume and lasts for as long as the breath of the tuner allows, but its accuracy often leaves much to be desired (depending on its quality, of course).

The present tuner may be considered as a very accurate electronic pitch-pipe. Accuracy of pitch is ensured by a quartz crystal, since even a mediocre crystal does not deviate from its nominal frequency by more than 100 p.p.m. (0.01%).

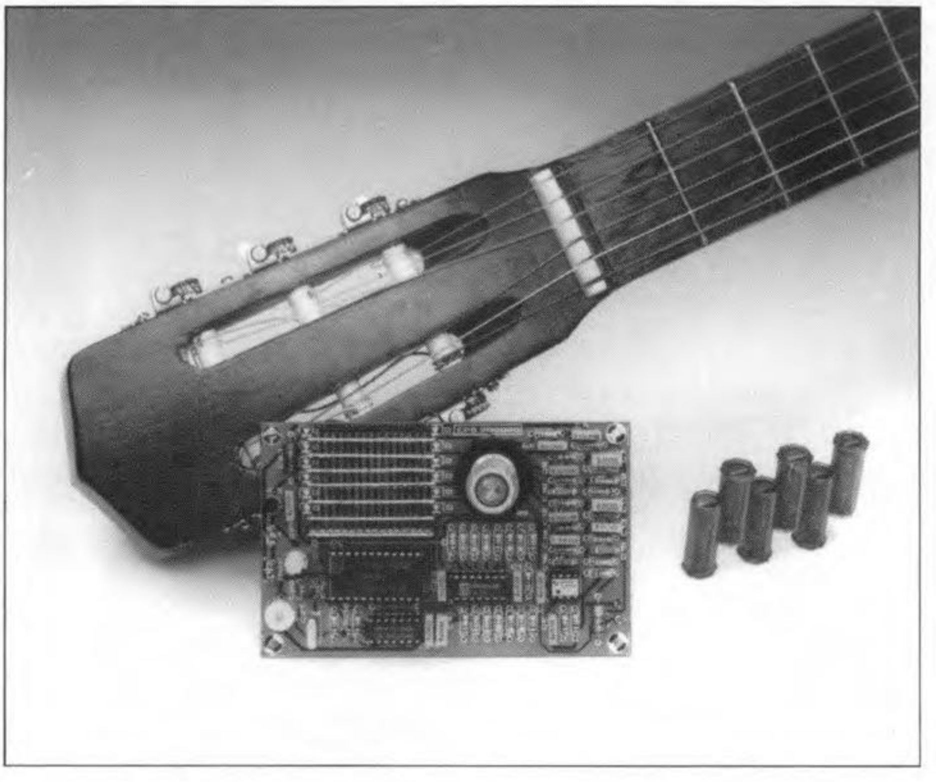

Fig. 1. General view of the electronic guitar tuner.

The six tones (US: notes)* required for tuning a guitar are obtained by dividing the crystal frequency with a presettable divider. Once the divider settings have been established, the six tones are always correct with respect to one another. Tuning is further simplified by the tone being a pure sine wave at constant loudness level. And, of course, the best acoustic measuring instrument available is used: the human ear. This has the additional advantage that the tuner's ear becomes trained in distinguishing between different tones: practice makes perfect.

Methods of tuning

Instruments may be tuned to the '**natural scale**' (which is deducible by physical laws); to **mean-tone temperament** (which gives a close approximation to natural tuning); and to **equal temperament** (in which musical intervals are moved away from the natural scale to fit them for practical performance). In equal temperament tuning, each semitone is made an equal interval. In other words, the twelve tones in an octave are equidistant on a logarithmic frequency scale, that is, each tone has a frequency that is $2^{1/12}$ (=1.05946) times greater than the preceding one. The advantage of this is that the instrument may be played in virtually any key. The disadvantage, however, is that the tones do not sound 'natural', which is especially

Table 1

		Harmonic tuning	Equal temp. tuning
$4f_{E2} = 3f_{A2} = 330.00$ Hz $\therefore f_{E2}$	=	82.50 Hz	82.41 Hz
$4f_{A2} = 3f_{D1} = 440.00$ Hz $\therefore f_{A2}$	=	110.00 Hz	110.00 Hz
$4f_{D1} = 3f_{G1} = 586.67$ Hz $\therefore f_{D1}$	=	146.67 Hz	146.83 Hz
$4f_{G1} = 3f_{B1} = 782.22$ Hz $\therefore f_{G1}$	=	195.56 Hz	196.00 Hz
$4f_{B1} = 3f_{A2} = 990.00$ Hz $\therefore f_{B1}$	=	247.50 Hz	246.94 Hz
$4f_{E2} = \quad f_{E}$	=	330.00 Hz	329.63 Hz

*In the United Kingdom, and most other English-speaking countries outside North America, a **tone** means a "musical sound consisting of a 'pure' note". In the USA, this is called a **note**. Similarly, "a single sound of given pitch and duration" is called a **note** in the UK and a **tone** in the USA.

no. of harmonic	1	2	3	4	5	6	7	8	9
length of string vibrating	entire	$\frac{1}{2}$	$\frac{1}{3}$	$\frac{1}{4}$	$\frac{1}{5}$	$\frac{1}{6}$	$\frac{1}{7}$	$\frac{1}{8}$	$\frac{1}{9}$

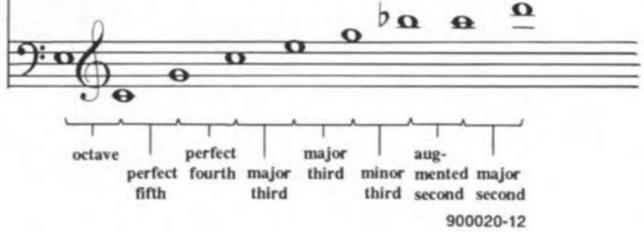

Fig. 2. Relation between harmonic frequencies and the length of string that is vibrating.

noticeable when concords are played.

In general, most guitarists prefer to tune their instrument to flageolet-notes, which means **harmonics**. The name derives from the supposed resemblance of these thin-sounding notes to those of the flageolet, an obsolete 6-holed wind instrument.

The name 'harmonic' is an abbreviation for 'harmonic tone', that is, one of the so-called harmonic series. The lowest of such tones, the 'fundamental', is called the 'first harmonic', the next lowest, the 'second harmonic', and so on. However, 'playing on harmonics' on a guitar really means 'harmonics without the first', since the fundamental is the 'normal' sound.

It is fairly simple to make the strings of a

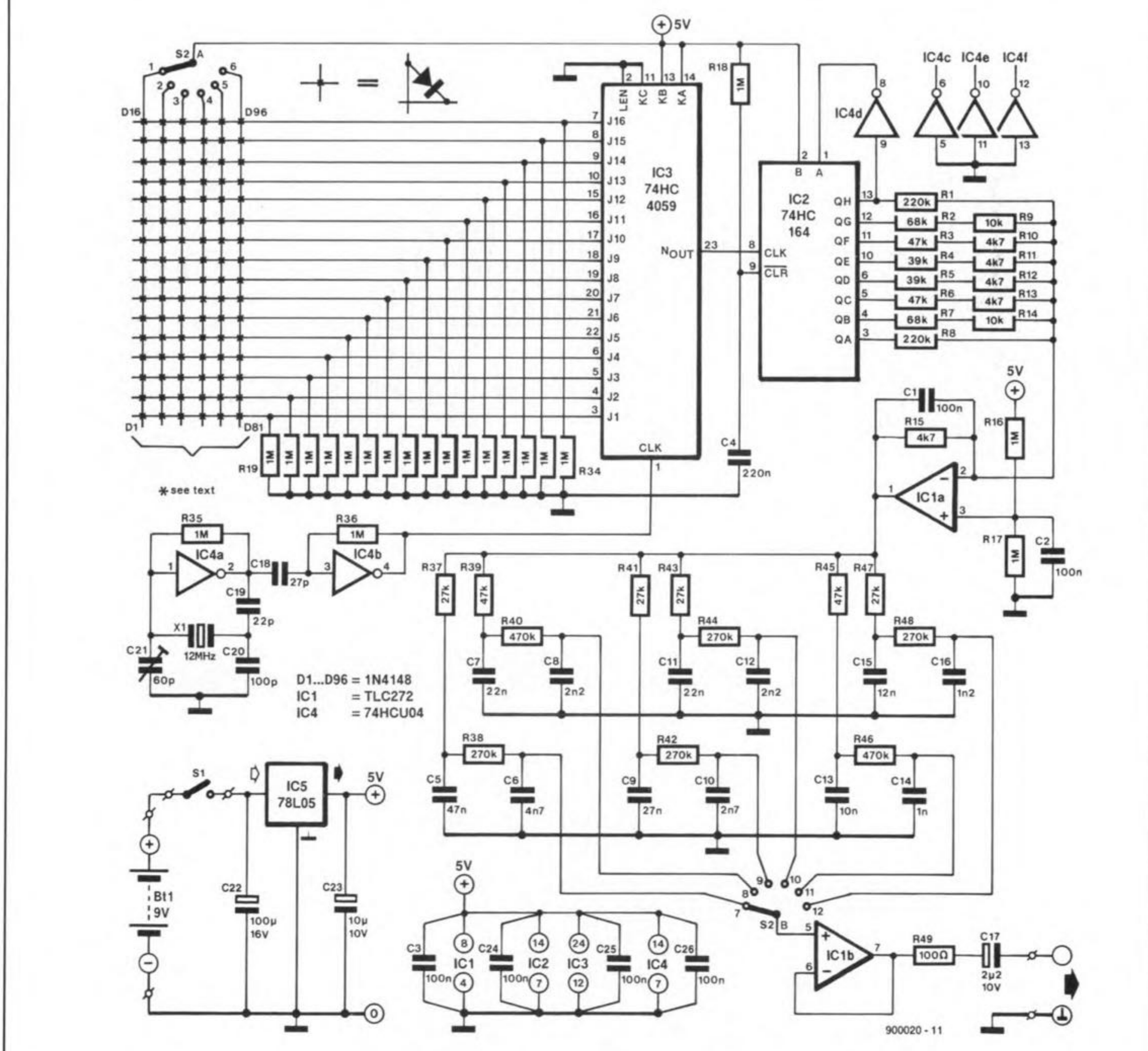

Fig. 3. Circuit diagram of the guitar tuner. Tuning frequencies are determined by the diode matrix D1–D81.

guitar vibrate at the second, third and fourth harmonic. This is done by setting the string vibrate not as a whole length but in fractional parts of its length. To obtain the second harmonic, the string is vibrated over half its length, for the third harmonic over a third of its length, and so on (see Fig. 2).

On a correctly tuned guitar, equal harmonic frequencies can be found on two different strings. For instance, the fourth harmonic of the lower E-string has the same frequency as the third harmonic of the A-string. In technical terms, $4f_{E2} = 3f_{A2}$. Relations between harmonics on other strings are shown in Table 1; if these are as shown, the guitar is tuned correctly. The frequencies are based on the International Concert Pitch, A = 440 Hz, whose second lower harmonic is A_2 = 110 Hz. From this, the other frequencies can be calculated. Note that the subscript or exponent of the tone following the string name indicates the number of octaves the tone is above or below Middle C respectively. The number 0 is traditionally omitted.

For completeness' sake, the table also shows the frequencies when the guitar is tuned to equal temperament. The choice is yours! Equal temperament is normally preferred for playing in a group, but for solo playing most guitarists choose harmonic tuning because that gives a 'smoother' sound. See also Table 2.

The tuner

The design of the tuner is far simpler than the theory behind the tuning. Its circuit, see diagram in Fig. 3, may be divided into four, excluding the power supply: a crystal oscillator, IC4a; a presettable frequency divider, IC3; a sine-wave shaper, IC2a-IC1a; and a six-stage *RC* output filter.

The oscillator is (and must be) an unbuffered inverter, and IC4a is therefore an HCU type. If you have a frequency counter, adjust C21 to give an oscillator frequency of exactly 12 MHz; otherwise, just set the trimmer to the centre of its travel. The oscillator is coupled to the clock (CLK) input of the divider.

The divider has four groups of four BCD (binary-coded decimal) inputs, J1–J16, with each of which one digit of the divisor, *k*, is set (max. 9999). Group J1–J4 sets the units; J5–J8 the tens; J9–J12 the hundreds; and J13–J16 the thousands. The setting is accomplished with the aid of a diode matrix, D1–D96 (not all of which are required).

The presence or absence of diodes determines the divisor for each of the six tones selected with S2. The presence of a diode causes a logic 1, and the absence a logic 0, at the relevant J-input of IC3. See also Table 2.

The output signal of the divider has a fre-

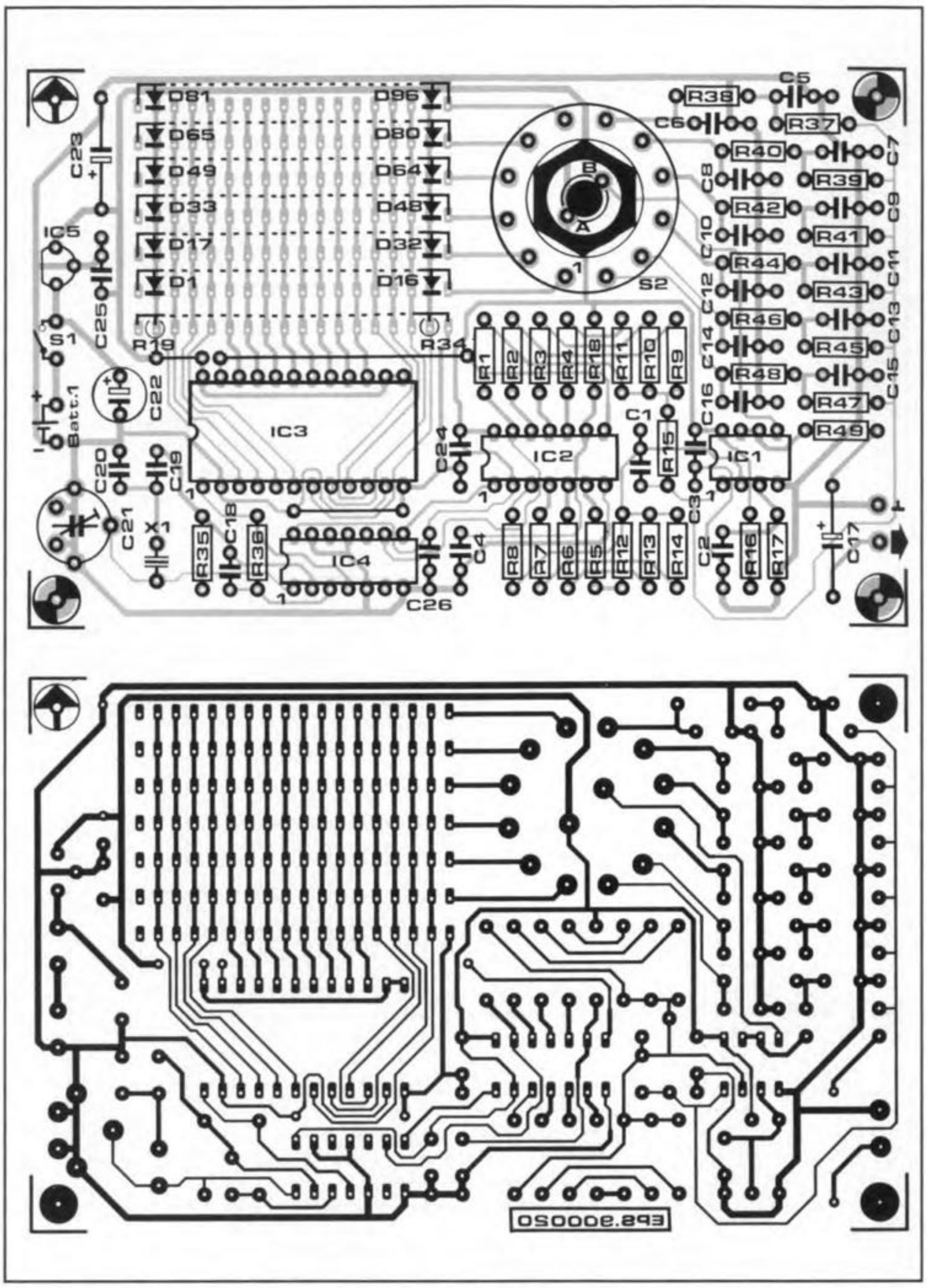

Fig. 4. The printed-circuit board for the tuner is single-sided to allow the diode matrix and the pull-up resistors to be mounted upright.

PARTS LIST

Resistors:
R1, R8 = 220 k
R2, R7 = 68 k
R3, R6, R39, R45 = 47 k
R4, R5 = 39 k
R9, R14 = 10 k
R10–R13, R15 = 4k7
R16–R36 = 1 M
R37, R41, R43, R47 = 27 k
R38, R42, R44, R48 = 270 k
R40, R46 = 470 k
R49 = 100 Ω

Capacitors:
C1,–C3, C24–C26 = 100 n
C4 = 200 n
C5 = 47 n
C6 = 4n7
C7, C11 = 22 n
C8, C12 = 2n2
C9 = 27 n
C10 = 2n7
C13 = 10 n
C14 = 1 n
C15 = 12 n
C16 = 1n2
C17 = 2µ2, 10 V, axial
C18 = 27 p
C19 = 22 p
C20 = 100 p
C21 = trimmer, 60 p
C22 = 100 µF, 16 V, radial
C23 = 10 µF, 10 V, axial

Semiconductors:
D1–D96 = 1N4148 (number required depends on tuning - see Table 2)
IC1 = TLC272
IC2 = 74HC164
IC3 = 74HC164
IC4 = 74HCU04
IC5 = 78L05

Miscellaneous:
S1 = single-pole switch
S2 = 2-pole, 6-position rotary switch for PCB mounting
X1 = 12 MHz quartz crystal
Bt1 = 9-V (PP3) battery and associated clip.
Enclosure, preferably ABS, 190×100×28 mm(approx.)
PCB Type 900020

quency that is 16 times higher than required. Clocked by this signal, the sine-wave shaper produces a pure sinusoidal signal at the right frequency at the output, pin 1, of IC1a.

Circuit IC2a is an eight-bit shift register that has been arranged to accept logic 1s when QH is low and logic 0s when QH is high. A reset at switch-on, provided by R18-C4, ensures that at the onset of operation always eight 0s are input first to the QA–QH inputs of the divider and then eight 1s.

Table 2

S2 posn.	Tone	Frequency	Diodes to be used	Divisor
		(Equal temperament tuning)		
1	f_{E2}	82.41 Hz	1; 9; 13; 16	9101
2	f_{A2}	110.00 Hz	20; 21; 28; 30; 31	6818
3	f_{D1}	146.83 Hz	36; 41; 45; 47	5108
4	f_{G1}	196.00 Hz	49; 50; 51; 54; 60; 61; 62	3827
5	f_{B1}	246.94 Hz	65; 66; 67; 69; 70; 77; 78	3037
6	f_{E}	329.63 Hz	81; 83; 85; 86; 87; 90; 94	2275
		(Harmonic tuning)		
1	f_{E2}	82.50 Hz	5; 8; 13; 16	9090
2	f_{A2}	110.00 Hz	20; 21; 28; 30; 31	6818
3	f_{D1}	146.70 Hz	33; 34; 37; 41; 45; 47	5113
4	f_{G1}	195.60 Hz	49; 51; 53; 54; 60; 61; 62	3835
5	f_{B1}	247.50 Hz	69; 70; 77; 78	3030
6	f_{E}	330.00 Hz	81; 82; 85; 86; 87; 90; 94	2273

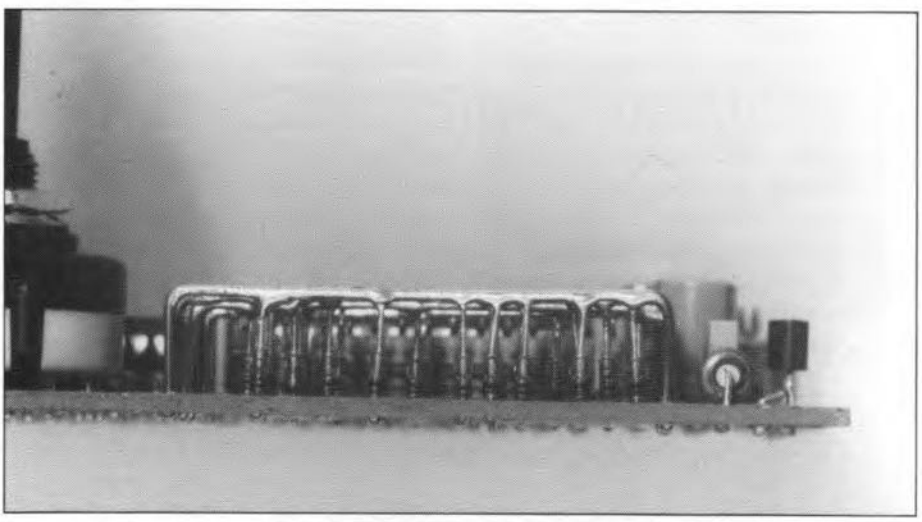

Fig. 5. All diodes used and the pull-up resistors must be mounted upright between the board and a bridge of stout circuit wire.

Fig. 6. Suggested front panel layout for the tuner.

The logic levels at the output of IC2 are translated into a resistance that, in conjunction with R15, determines the gain of IC1a.

The shape of the output of IC1a is nearly sinusoidal, mainly because of C1 in the feedback loop. Nevertheless, the 15th and 17th harmonics are still fairly strong but, since these are, in frequency, a good way from the fundamental, a simple second-order *RC* filter is sufficient to suppress them. However, the distance (in frequency) between the fundamental and these harmonics is not so large as to make one filter suffice for all six tones: each of the tones requires a separate filter.

The six filters are switched into circuit by S2b. The waveform at the pole of this switch is a good sine wave that has less than 0.04% harmonic distortion. Even so, the signal is still buffered by IC1b. The cut-off frequency of the filters is about 60% of the frequency at the –3 dB point.

The output of the tuner, which is protected against short-circuits by R49, is suitable for connecting to a variety of amplifiers. Its level depends to some degree on the frequency and lies between 450 mV and 600 mV r.m.s. Because of this dependency on frequency, the loudness appears to remain more constant than when the output level is kept constant.

Construction and setting up

The tuner is best built on the printed-circuit board shown in Fig. 4 and then fitted in a suitable enclosure of about 190×100×28 mm. Mount all pull-down resistors and the required diodes (34 for equal temperament tuning and 33 for harmonic tuning – see Table 2) upright as shown in Fig. 5.

The divisors are obtained as briefly explained earlier. For instance, the divisor, *k*, for frequency f_{G1} (equal temperament tuning) is obtained by placing diodes D49, D50 and D51 on to J1, J2, and J3 respectively and no diode to J4 to give binary number 0111 = decimal 7; diode D54 on to J6 and no diodes to J5, J7 and J8 to give binary number 0010 = decimal 2; diode D60 on to J12 and no diodes to J9, J10 and J11 to give binary number 1000 = decimal 8; and diodes D61 and D62 onto J13 and J14, and no diodes to J15 and J16 to give binary number 0011 = decimal 3. The divisor is thus 3827.

The divisor may be calculated from $k = 12\times10^6 / 16f$. If frequencies different from those in this article are used, bear in mind that the filters must be adapted accordingly.

The most practical power supply is a 9-V (PP3) battery. The average operational current drain is 12 mA. The battery will last a long time as shown by the prototype still working satisfactorily when the battery voltage had dropped to 4 V. ■

guitar vibrate at the second, third and fourth harmonic. This is done by setting the string vibrate not as a whole length but in fractional parts of its length. To obtain the second harmonic, the string is vibrated over half its length, for the third harmonic over a third of its length, and so on (see Fig. 2).

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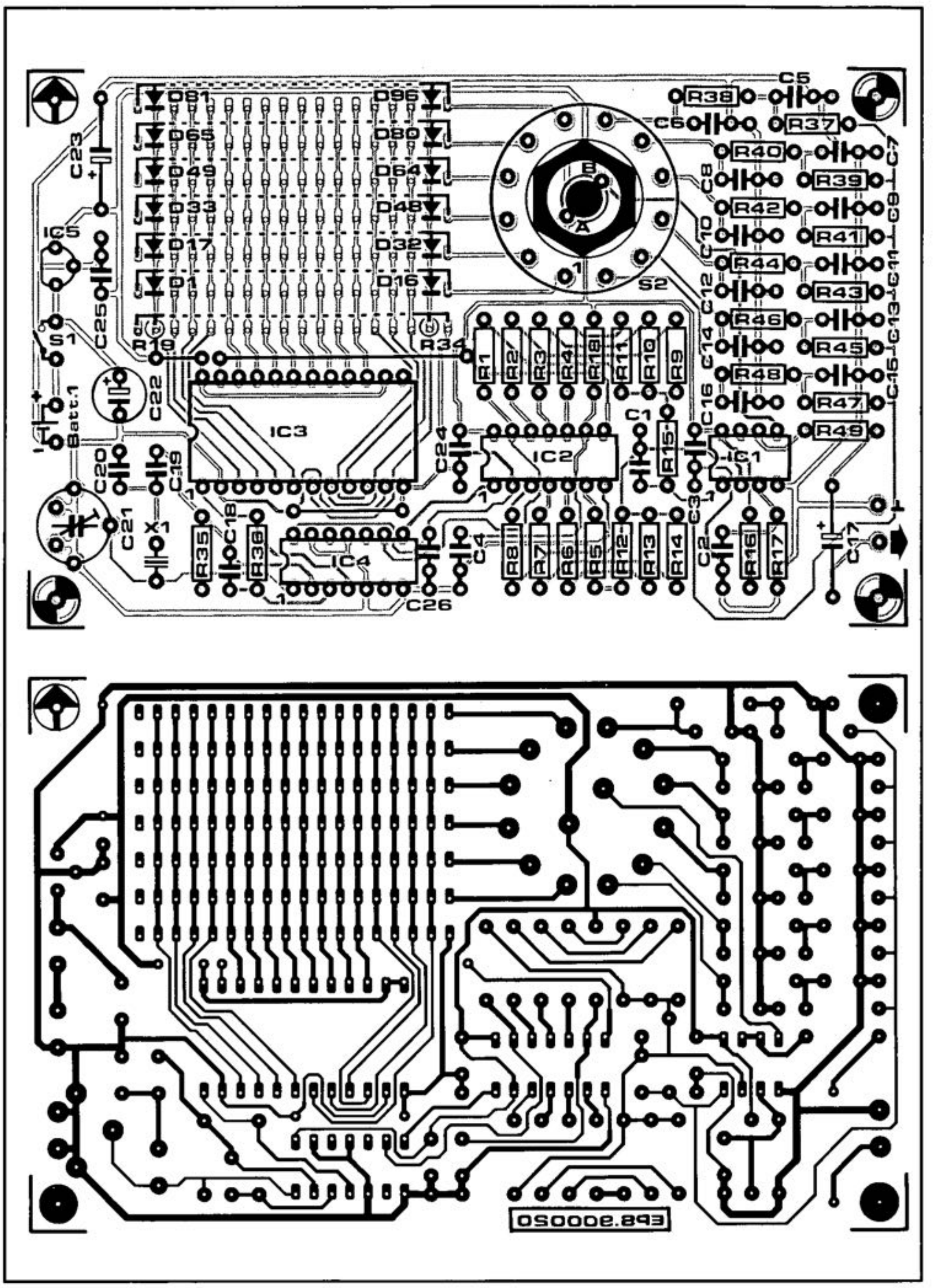

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C5 = 47 n
C6 = 4n7
C7, C11 = 22 n
C8, C12 = 2n2
C9 = 27 n
C10 = 2n7
C13 = 10 n
C14 = 1 n
C15 = 12 n
C16 = 1n2
C17 = 2μ2, 10 V, axial
C18 = 27 p
C19 = 22 p
C20 = 100 p
C21 = trimmer, 60 p
C22 = 100 μF, 16 V, radial
C23 = 10 μF, 10 V, axial

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X1 = 12 MHz quartz crystal
Bt1 = 9-V (PP3) battery and associated clip.
Enclosure, preferably ABS, 190×100×28 mm(approx.)
PCB Type 900020

IN QUEST OF A PANGRAM – PART 3

V g00103 - III

by Lee C.F. Sallows

Logological space

Though a bitter disappointment, the failure of the quest was not yet an irreversible defeat. A remote chance lingered that the magic combination lay yet undetected just outside the range of number-words examined. More promisingly, alternative translations remained to be explored. At the top of the list was "this pangram comprises ...", a rendering of the Dutch *bevat* on a par with "contains". This would only entail a new set of initial text constants.

The prospect of a further month in purgatory, however, was anything but inviting. Yet, much had happened during the long weeks of waiting. In the range-limiting stratagem used to shorten the previous run had lain the seed of a powerful new development. Many hours' thought had been given to this, and already detailed preparations were in hand for a Mark II version of the machine incorporating extensive modifications.

Consider the number-words in the range for *y*: *three*, *four*, *five*: the letter *y* itself occurs in none of them. Put differently, whichever of *y*'s PROFILES may be activated, the actual number of *y*'s can never be affected; in this sense, *y* is an independent variable. Great advantage can be taken of this by adding new circuitry which *measures* the number of *y*'s present in the currently activated combination and uses the result to switch in the appropriate *y*-PROFILE. In short, the *y*-counter can be replaced by an *automatic number-word selector*. And discarding the *y*-counter from the cascade will mean *dividing the running time by three* (see Fig. 5).

The real power of this refinement emerges on seeing that the same trick can be worked for any letter not appearing in the number-words making up its own range. *G* and *l* are two such; provided *six* is dropped from its range, so is *x*. This then was the scheme to be realized in the blueprint for the new Mark II machine. With the *g*, *l*, *x*, *y* counters removed from the cascade, running time falls to only

$(8\times6\times1\times6\times7\times1\times7\times6\times6\times7\times7\times6\times6\times7\times1\times1)/10^6$

seconds or *one hundred and five minutes*! The perspective opened up by this dramatic improvement carried further implications in its wake.

With the ability to explore so quickly, means would be required for loading of different initial text constants. Though electrically trivial, a flexible resistance-selection method was difficult to implement in the machine as it stood. The final (and not altogether satisfactory) system chosen uses a set of four tiny switches for each channel. The latter works in binary fashion, so that a constant or "weight" of anything from 0 through 15 letters can be introduced. Incorporating this bank of 16×4 PRESET LETTER WEIGHTS switches on the front panel (see photo below) involved some major surgery to the machine.

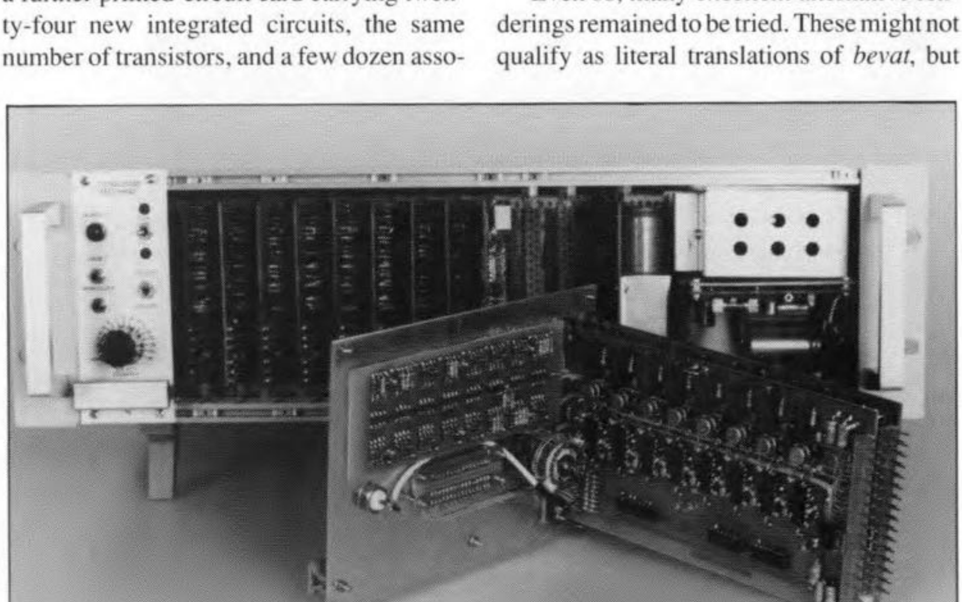
Another benefit of ultra-fast logological space travel is the chance to prospect further afield, that is, to expand ranges. Even if all twelve remaining counters are allocated a range length of 8 (the maximum available in this machine), running time comes out to only $8^{12}/10^6$ seconds = 19.08 hours. In two cases, *l* and *y*, the ranges of auto-selected letters may themselves be increased, an expansion that has its uses with initital texts containing *l*'s and *y*'s; for instance, "This pangram employs ...". The *g* in *eight* and *x* in *six* make further extension impossible for *g* and *x*. In reality, impatience to get on dissuaded me from expanding range lengths until later so that running time was kept below two hours during initial explorations.

Besides serious mechanical alterations, the modifications sketched above called for a further printed-circuit card carrying twenty-four new integrated circuits, the same number of transistors, and a few dozen associated components. The increased electrical drain meant in turn an extra d.c. power supply. Space was cramped, and the rise in internal heat dissipation threatened to upset the temperature-sensitive differential amplifiers. Notwithstanding these demands and difficulties, within a month the new souped-up Pangram Machine Mark II stood poised for its maiden flight.

Following a last-minute test with the modified initial text constants, now easy to enter via the front-panel switches, I started off with a recheck of "This pangram contains ...". With running time down to under two hours, one could afford to be thorough. This time there was no wine, no ceremony, no Velasquez and, as anticipated, no result.

In the mean time I'd worked out the initial text constants for "This pangram *comprises* ...", and as soon as the first run was over, I loaded these and set the machine searching again. Two hours later, the counter LEDS showed that the second run had been completed, and I was confronting another disappointment. That truly was a tragedy: it meant that no really perfect translation of the Dutch pangram existed. It seemed to me an unwarranted injustice: and, brushing aside a tear, I marked it down as another of the things I mean to ask God about on Judgement Day.

Even so, many excellent alternative renderings remained to be tried. These might not qualify as literal translations of *bevat*, but



Two printed-circuit cards behind the right-hand front panel carry the sixteen BALANCE-detectors. The circular metal cans are differential amplifiers, below which a line of eight integrated circuits form the ZERO? detectors. On the panel itself are the initial text-constant resistors selected by means of the PRESET LETTER WEIGHTS switches.

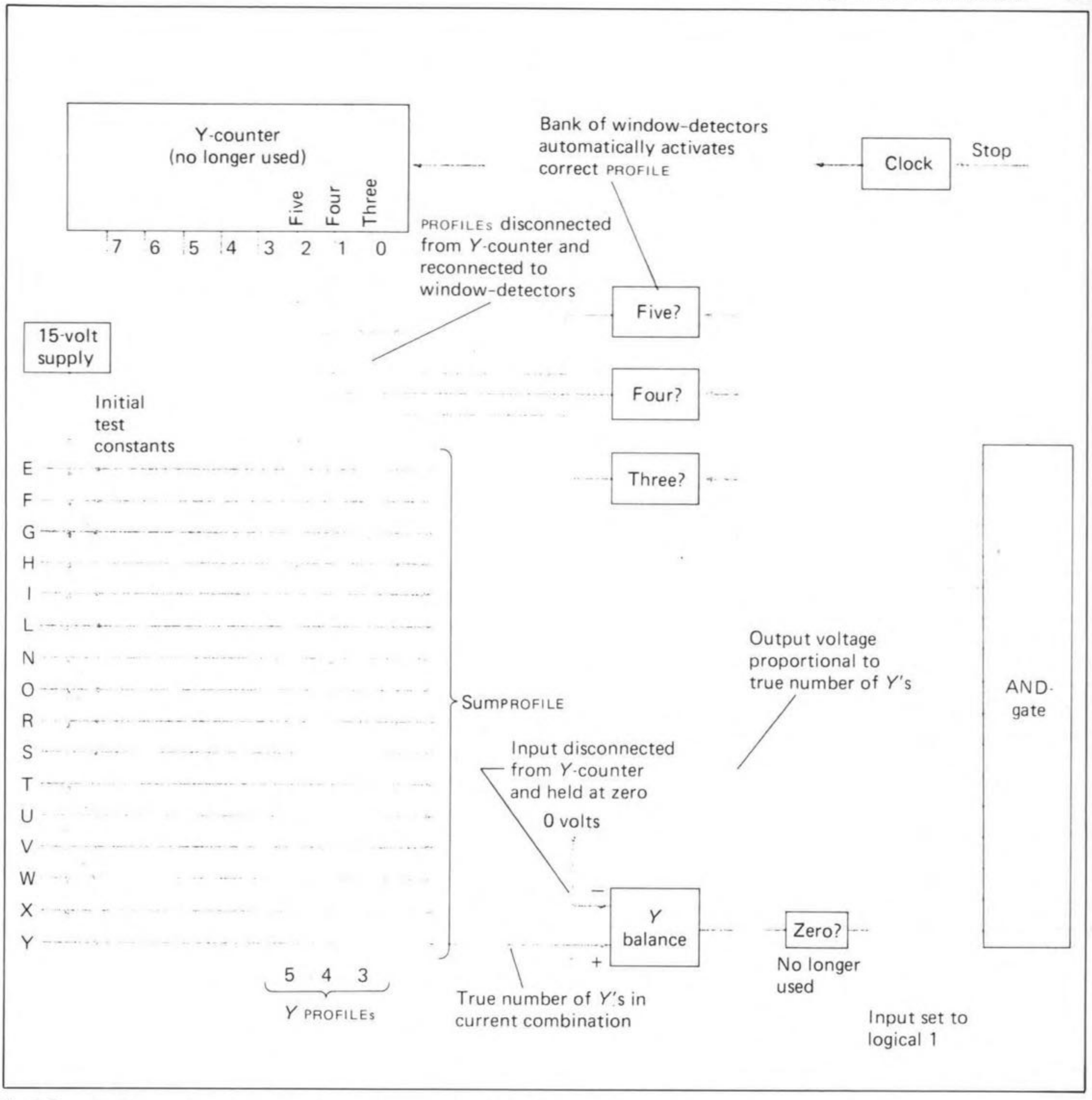


Fig. 5. Example of automatic number-word selection applied to the letter *y*. A voltage proportional to the number of *y*'s occurring in the present combination is classified by a bank of three window-detectors, one of whose outputs will activate the appropriate PROFILE.

would at least preserve the spirit of the original. "This pangram *comprises*..." was therefore followed in quick succession by "This pangram *consists of*", "*is composed of*", "*uses*", "*employs*", and "*has*". Everyone of them without success!

By now I was beginning to wonder just how long this might go on. Given a random introductory text of, say, twenty-five letters, what is the probability that an associated self-enumerating list exists? Short of examining all possible twenty-five letter strings one at a time, I saw no way of answering the question. One in a hundred? One in a million? As it happens, the answer turns out to be something closer to one in ten.

On the second day of exploration I was sitting in front of the machine during its eighth run when suddenly the EUREKA!-lamp came on and my stomach turned a somersault. Rigid with excitement, I carefully decoded the LED displays into the set of number-words represented. A painstaking check completely verified the following perfect pangram:

This pangram lists four *a*'s, one *b*, one *c*, two *d*'s, twenty-nine *e*'s, eight *f*'s, three *g*'s, five *h*'s, eleven *i*'s, one *j*, one *k*, three *l*'s, two *m*'s, twenty-two *n*'s, fifteen *o*'s, two *p*'s, one *q*, seven *r*'s, twenty-six *s*'s, nineteen *t*'s, four *u*'s, five *v*'s, nine *w*'s, two *x*'s, four *y*'s, and one *z*.

I leave it to my readers to imagine the scenes of wild intemperance following upon this victory. Despite a hangover, next morning copies of the pangram were happily handed out among friends and colleagues who had patiently borne with me through the long months of pangrammania. Notable, if unsurprising, was that nobody felt disposed to examine the sentence for a discrepancy. Not unnaturally, I came in for a few kind words of congratulation, and some even looked at me with an unspoken: "How does it feel to climb Everest?" on their lips. Like a dishrag, actually; I still hadn't recovered from the previous evening's celebrations.

The zenith of glory was yet to come. Returning home at lunch-time, I found a magnificent trophy awaiting. I had set the machine running once more, early in the morning, and it had halted again at a new solution. Changing "and" to "&" in the natural English rendering of Rudy Kousbroek's pangram, a last desperate bid for a perfect magic translation had finally met with success. The Quest for the Pangram had ended in triumph!

This pangram contains four *a*'s, one *b*, two *c*'s, one *d*, thirty *e*'s, six *f*'s, five *g*'s, seven *h*'s, eleven *i*'s, one *j*, one *k*, two *l*'s, two *m*'s, eighteen *n*'s, fifteen *o*'s, two *p*'s, one *q*, five *r*'s, twenty-seven *s*'s, eighteen *t*'s, two *u*'s, seven *v*'s, eight *w*'s, two *x*'s, three *y*'s, & one *z*.

More and more pangrams

Looking back on it, I suppose the failure of the Mark I machine to find the pangram was a piece of good fortune. I mean, otherwise, the fast and flexibile research instrument realized in the Mark II model may never have come into being. As it was, I could now experiment at will, initially confined only to the spectrum of possibilities defined by the given set of number-word ranges. This was an important limitation, since pangram-oriented ranges are unlikely to prove fertile in canvassing for autograms in general. In a self-enumerating *pangram*, the non-critical letters *a*, *b*, *c*, *d*, *j*, *k*, *m*, *p*, *q*, and *z* are likely to be prefixed by the words *one* or *two*; the frequency of *o*'s, *n*'s, *e*'s, *t*'s, and *w*'s is thereby significantly slanted. Save in special cases, non-pangrams would give rise to distributions lying outside the scope of the machine.

The exploration I now embarked upon was a source of great fun and interest. A thoughtful Platonist can only wonder at some of the eternal Truths that God has seen fit to leave scattered about in the regions traversed by the machine. An early find was a somewhat wry specimen I couldn't resist sending off to Rudy Kousbroek. I suppose it might be described as a dead-pan-gram.

This pungram boasts four *a*'s, two *b*'s, one *c*, two *d*'s, twenty-eight *e*'s, seven *f*'s, three *g*'s, five *h*'s, nine *i*'s, one *j*, one *k*, one *l*, two *m*'s, twenty *n*'s, fifteen *o*'s, two *p*'s, one *q*, five *r*'s, twenty-seven *s*'s, twenty-one *t*'s, three *u*'s, six *v*'s, ten *w*'s, two *x*'s, five *y*'s, and one *z*.

Doubtless he will find little difficulty in producing a magic Dutch translation of this sentence. Another example which seemed worth drawing to his attention was:

This pangram containeth five *a*'s, one *b*, two *c*'s, two *d*'s, twenty-five *e*'s, seven *f*'s, two *g*'s, four *h*'s, ten *i*'s, one *j*, one *k*, one *l*, two *m*'s, twenty *n*'s, sixteen *o*'s, two *p*'s, one *q*, five *r*'s, twenty-six *s*'s, twenty-one *t*'s, three *u*'s, six *v*'s, ten *w*'s, four *x*'s, five *y*'s, and one *z*.

I don't know whether he believed my tale of it having turned up among the marginalia in a folio edition of *Macbeth*. Probably not. The Dutch have never entirely succeeded in shaking off the legacy of German Scepticism.

If the above squibs suggest frivolity, it must be put down to the sudden release of tension after months of unrelenting effort. To have sought so long and hard for a single jewel only to end up with a (potential) embarrassment of riches was an unhinging experience. For a while I reconnoitred without any clear plan. Among other diversions, sentences incorporating names of friends provided entertainment. It was interesting to find how readily some of these lent themselves to immortality:

This pangram for Doug Hofstadter* contains five *a*'s, one *b*, two *c*'s, three *d*'s, twenty-seven *e*'s, seven *f*'s, three *g*'s, six *h*'s, ten *i*'s, one *j*, one *k*, one *l*, two *m*'s, twenty *n*'s, sixteen *o*'s, two *p*'s, one *q*, nine *r*'s, thirty *s*'s, twenty *t*'s, four *u*'s, six *v*'s, seven *w*'s, four *x*'s, five *y*'s, & one *z*.

In this way, many pangrams were unearthed, and the data derived from them shed new light on the relation between initial text values and the ranges in which solutions could be expected. This information could be plugged back into the machine through altering ranges so as to maximize the probability of future success with certain texts. After a time, the facility achieved in prospecting for nuggets prompted an ambitious new research programme.

The immortal lie

This sentence doesn't contain two *a*'s, three *c*'s, two *d*'s, twenty-six *e*'s, six *f*'s, three *g*'s, seven *h*'s, eight *i*'s, fifteen *n*'s, ten *o*'s, seven *r*'s, twenty-eight *s*'s, twenty-one *t*'s, four *u*'s, four *v*'s, seven *w*'s, three *x*'s, four *y*'s & two *&*'s.

A shortcoming of logology, I find, is its absence of underlying structure. Like mathematics, it manifests itself in precisely defined chains of atomic symbols, yet lacks the intrinsic patterning, the symmetry of the former. Anyone with a feeling for mathematical form will probably regret this deficit, too. Autograms, however, embody a peculiar fusion of both fields, an improbable marriage of arbitrary convention with arithmetical necessity. The unexpected possibilities they point to re-echo mathematical affinities. In particular, among other high-order entities now appearing over the horizon of this strange realm are the counterparts of *numerical series*. The most obvious of these now became the focus of machine investigation:

This first pangram has five *a*'s, one *b*, one *c*, two *d*'s, twenty-nine *e*'s, six *f*'s, four *g*'s, eight *h*'s, twelve *i*'s, one *j*, one *k*, three *l*'s, two *m*'s, nineteen *n*'s, twelve *o*'s, two *p*'s, one *q*, eight *r*'s, twenty-six *s*'s, twenty *t*'s, three *u*'s, five *v*'s, nine *w*'s, three *x*'s, four *y*'s, and one *z*.

This second pangram totals five *a*'s, one *b*, two *c*'s, three *d*'s, twenty-nine *e*'s, six *f*'s, four *g*'s, seven *h*'s, ten *i*'s, one *j*, one *k*, two *l*'s, two *m*'s, twenty-one *n*'s, sixteen *o*'s, two *p*'s, one *q*, eight *r*'s, twenty-eight *s*'s, twenty-three *t*'s, four *u*'s, four *v*'s, nine *w*'s, three *x*'s, five *y*'s, and one *z*.

This third pangram contains five *a*'s, one *b*, two *c*'s, three *d*'s, twenty-six *e*'s, six *f*'s, two *g*'s, four *h*'s, ten *i*'s, one *j*, one *k*, two *l*'s, two *m*'s, twenty-two *n*'s, seventeen *o*'s, two *p*'s, one *q*, seven *r*'s, twenty-nine *s*'s, twenty-one *t*'s, four *u*'s, six *v*'s, eleven *w*'s, four *x*'s, five *y*'s, and one *z*.

Prolongation of the series, written out in full, would be too space-consuming. Figure 6 presents an abbreviated record of the first twenty-five terms, with figures standing for number-words. Note the use of a distinct verb in each case. This is not always necessitated, since the same word combined with different ordinals may also generate solutions. The employment of a different verb each time seemed to me demanded on esthetic grounds.

The uncovering of this series is, in my opinion, among the most felicitous results of the machine. Though a mere matter of patient search, hundreds of running hours were involved. In one case, more than forty verbs were tried before locating a solution. On the average, though, winning combinations can be found for one in eight initial texts. This figure is empirically derived, of course. It seems to me worth pondering that (to my knowledge) no existent mathematical technique is able to assign even a rough value to the probability of detecting a solution. Conceivably, artificial languages or, at least, artificially constructed number-word systems might be of use in gaining further insight into this.

The list published here is not as long as I could have made it. Eventually, I hope, one

* American physicist, born 1945. Well-known for his many writings, among which *Gödel, Escher, Bach: an Eternal Golden Braid*, which won the 1980 Pulitzer Prize for general non-fiction. [Ed]

Representation of 25 Pangrams

This N^{th} pangram	-------	A	B	C	D	E	F	G	H	I	J	K	L	M	N	O	P	Q	R	S	T	U	V	W	X	Y	Z
1st	has	5	1	1	2	29	6	4	8	12	1	1	3	2	19	12	2	1	8	26	20	3	5	9	3	4	1
2nd	totals	5	1	2	3	29	6	4	7	10	1	1	2	2	21	16	2	1	8	28	23	4	4	9	3	5	1
3rd	contains	5	1	2	3	26	6	2	4	10	1	1	2	2	22	17	2	1	7	29	21	4	6	11	4	5	1
4th	numbers	4	2	1	2	29	7	2	6	10	1	1	1	3	23	14	2	1	9	26	20	5	5	9	3	5	1
5th	embraces	5	2	2	2	29	7	3	8	10	1	1	1	3	20	13	2	1	9	26	24	3	4	10	2	5	1
6th	harbours	5	2	1	2	28	7	4	7	10	1	1	1	2	21	15	2	1	7	28	20	4	6	9	3	5	1
7th	counts	4	1	2	2	30	5	3	7	9	1	1	1	2	23	16	2	1	7	28	21	4	7	9	2	5	1
8th	tallies	5	1	1	2	30	5	3	7	10	1	1	3	2	20	14	2	1	6	27	21	2	7	9	2	5	1
9th	exploits	4	1	1	2	28	7	4	8	13	1	1	2	2	22	16	3	1	9	26	23	5	4	9	4	5	1
10th	features	5	1	1	2	28	8	5	6	12	1	1	2	2	18	14	2	1	6	27	20	5	6	9	4	4	1
11th	utilizes	4	1	1	2	31	7	4	7	11	1	1	4	2	20	15	2	1	8	29	18	6	6	7	3	4	2
12th	tables	5	2	1	2	26	6	2	6	11	1	1	4	2	17	13	2	1	7	30	20	3	6	9	5	4	1
13th	includes	4	1	2	3	29	8	4	8	12	1	1	3	2	20	14	2	1	9	25	24	6	5	10	2	5	1
14th	recruits	4	1	2	2	28	8	4	7	10	1	1	1	2	20	15	2	1	10	26	24	6	3	9	3	5	1
15th	uses	4	1	1	2	30	7	2	5	9	1	1	1	2	22	16	2	1	5	27	21	3	7	10	2	5	1
16th	subsumes	4	2	1	2	30	7	4	8	10	1	1	1	3	21	15	2	1	8	29	21	6	4	7	3	5	1
17th	tabulates	6	2	1	2	28	7	3	5	10	1	1	2	2	20	14	2	1	6	29	24	5	6	10	4	5	1
18th	manifests	5	1	1	2	35	8	5	10	13	1	1	1	3	21	14	2	1	8	26	24	3	7	7	2	5	1
19th	assembles	5	2	1	2	35	6	5	10	12	1	1	4	3	18	12	2	1	8	28	23	3	7	9	2	4	1
20th	summons	4	1	1	2	29	7	3	5	11	1	1	2	4	22	16	2	1	6	28	21	5	6	10	4	5	1
21st	shows	4	1	1	2	29	6	3	6	11	1	1	3	2	22	16	2	1	8	29	21	4	4	11	5	6	1
22nd	displays	5	1	2	4	33	5	3	9	12	1	1	3	2	21	13	3	1	9	28	25	2	6	10	2	7	1
23rd	produces	4	1	2	4	26	6	2	4	10	1	1	2	2	22	17	3	1	9	29	21	6	4	11	5	6	1
24th	evinces	4	1	2	2	26	6	2	4	9	1	1	2	2	21	17	2	1	7	30	20	5	7	11	4	6	1
25th	discloses	4	1	2	3	32	7	3	9	11	1	1	3	2	20	14	2	1	9	28	25	3	5	10	2	6	1

Fig. 6. A representation of 25 pangrams. In the actual pangrams, the numbers in the first column would be replaced by "first", "second", ...,"twenty-fifth". The numbers in the main body of the table would also be replaced by number-words. The fourth word of each pangram is shown in the second column.

hundred will be reached. In the mean time, I can't help wondering how the discovery will strike others. Who could have foreseen such a possibility? Once upon a time it had seemed daring to believe a single gem might exist. The finding of a (potentially infinite) cluster of matching stones by far exceeds my greediest imaginings.

As I went along, I made up some new plug-in matrix cards that use different resistor sets so as to cast a wider net, able to embrace certain kinds of non-pangram autogram:

This autogram contains two *a*'s, two *c*'s, two *d*'s, twenty-eight *e*'s, five *f*'s, three *g*'s, eight *h*'s, eleven *i*'s, three *l*'s, two *m*'s, thirteen *n*'s, nine *o*'s, two *p*'s, five *r*'s, twenty-five *s*'s, twenty-three *t*'s, six *v*'s, ten *w*'s, two *x*'s, five *y*'s, and one *z*.

The apparent elegance of these can sometimes be deceptive; closer scrutiny may reveal imperfections. For instance, oughtn't "one *z*" to be regarded as a redundant curlicue? Its inclusion is clearly a gratuitous addition to the preceding text. Romantics may gaze indulgently at such ornament, but purists will point out that its real function is to contribute an extra *o*, *n*, and *e* merely in order to make the sentence work. Appending number-words is just a cunning way of disguising text-doctoring. Perhaps those with a sneaking affection for the solitary *z* will find consolation in:

This sentence contains three *a*'s, three *c*'s, two *d*'s, twenty-six *e*'s, five *f*'s, three *g*'s, eight *h*'s, thirteen *i*'s, two *l*'s, sixteen *n*'s, nine *o*'s, six *r*'s, twenty-seven *s*'s, twenty-two *t*'s, two *u*'s, five *v*'s, eight *w*'s, four *x*'s, five *y*'s, and only one *z*.

Here, the inclusion of "only" legitimizes the addition of "one *z*" by "proving" it was premeditated. Even so, the choice of letter remains arbitrary: a *q* would have done just as well. Classicists, however, will reject all *q*'s (whether straight or curly) and rightly insist on the crisp parsimony of:

This sentence employs two *a*'s, two *c*'s, two *d*'s, twenty-six *e*'s, four *f*'s, two *g*'s, seven *h*'s, nine *i*'s, three *l*'s, two *m*'s, thirteen *n*'s, ten *o*'s, two *p*'s, six *r*'s, twenty-eight *s*'s, twenty-three *t*'s, two *u*'s, five *v*'s, eleven *w*'s, three *x*'s, and five *y*'s.

It is odd to reflect that the existence of this minimal form seems to vitiate the objection raised against the first version; "one *z*" may be redundant, but it couln't have been thrown in just to make the sentence work! Subtleties of this kind should be kept in mind when trying to assess the relative merits of different specimens. ■

The final part of this article, reproduced by kind permission of Springer-Verlag, Heidelberg and New York, will appear in the November issue of *Elektor Electronius*.

SPEED CONTROL FOR 3-PHASE MOTORS

by F. P. Zantis

Three-phase motors, particularly asynchronous types, are very reliable and consequently popular prime movers. A drawback of these engines is, however, their fixed speed of rotation which depends on the frequency of the applied voltage. This article describes means of varying that frequency and thus controlling the speed of the motor over a wide range.

Until not so long ago, the only way of varying the frequency of the voltage driving a three-phase motor was the use of a rotary converter . However, the advent of power semiconductors has made possible the development of static frequency changers that convert an alternating voltage of a given frequency into one of a different frequency. These devices may also change the number of phases, for instance, single-phase current into three-phase current.

Apart from their application with prime movers, such frequency changers are also very useful in standby and emergency power supplies and in the control circuit of oscillators. In these applications, the widely varying direct output voltage of a battery is converted to an alternating voltage of steady level and frequency.

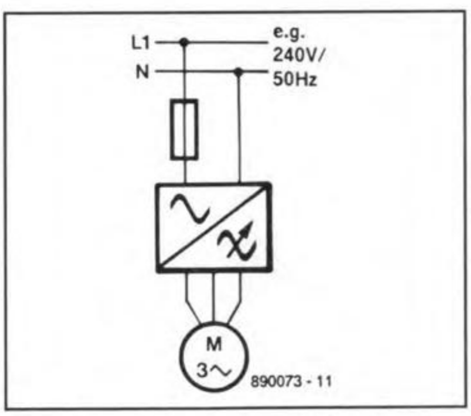


Fig. 1. Schematic representation of how a 3-phase motor may be operated from a single-phase supply under the control of a frequency changer.

A frequency changer to control the speed of a three-phase motor must have an output voltage whose level and frequency are both variable. A proportional change in the level and frequency of the voltage applied to the motor enables the speed of rotation of the motor to be varied at constant moment as is shown schematically in Fig. 1.

Speed control offers many advantages, such as a saving in energy, reduced maintenance costs, and optimalization or greater flexibility of operation.

Frequency changers

All frequency changers work on the same principle: the drive voltage (normally 240 V or 415 V, 50 Hz mains) is rectified, smoothed (filtered) and applied to the motor via an inverter (= dc-to-ac converter) as shown schematically in Fig. 2.

The construction of the various sections depends on the type of frequency changer. For instance, the power section of a small frequency changer is shown (greatly simplified) in Fig.3.

The **rectifier** is normally a conventional bridge type. The voltage source is loaded by the required current only, independent of the motor rating. That results in a good power factor which is virtually constant (typically, $\cos \phi = 0.97$) over the entire load and speed ranges.

The **smoothing circuit** is normally an *LC* low-pass filter, whose inductance protects the mains against transients. The large values of capacitance required with high loads are obtained by series and parallel connection of a number of high-voltage electrolytic capacitors as shown in Fig. 4. Matched resistors ensure correct division of the voltage.

The **inverter** consists of three pairs of transistors that are arranged in a star configuration. The three motor phases are connected cyclically at 120° intervals to, respectively, the positive and negative terminal of the smoothing filter, which results in a rotating field being induced in the motor. Appropriate control of the inverter enables the smooth, precise control of the output frequency. In general, three-phase motors may be operated at up to twice their rated speed. This means that, for example, a four-pole

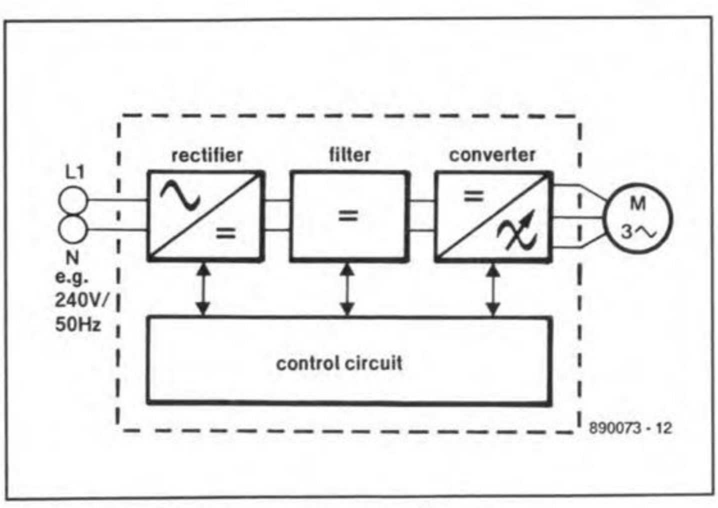


Fig. 2. Block diagram of a basic frequency changer.

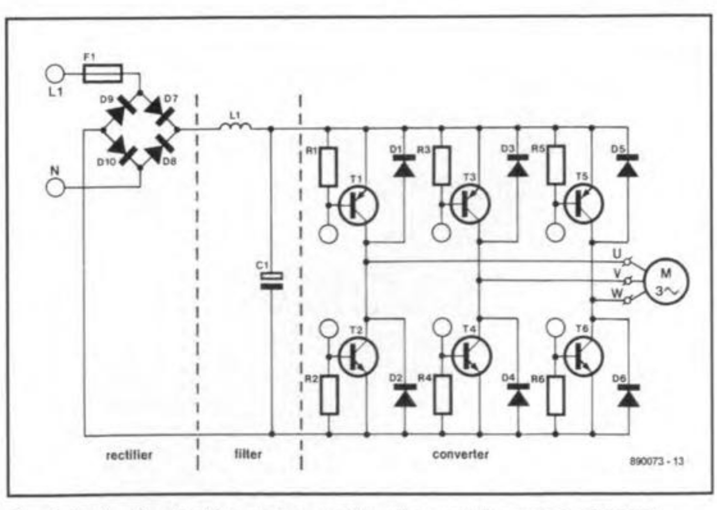


Fig. 3. Basic circuit of the power section of a small frequency changer.

three-phase motor may be operated at up to 3000 rev/min. Such control can not, however, be achieved by varying only the frequency.

To obtain a constant moment, the magnetic flux in the stator of the motor must be related to the set frequency. To that end, the motor voltage must be increased or reduced, as the case may be, in direct proportion to the frequency. When the frequency is higher than the rated frequency of the motor (50 Hz), the voltage can not rise, since the frequency changer can not generate a potential that is higher than the applied voltage.

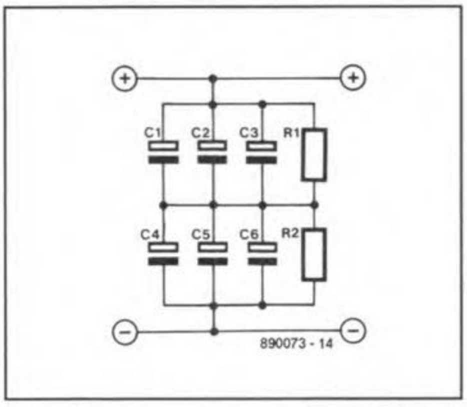


Fig. 4. The filter capacitance may consist of series and parallel arrangements of small capacitors.

This results in a reduction of the magnetic flux and thus of the moment. Operation above the rated frequency is thus not possible at the rated moment.

The basic relation between supply frequency, f_s, the ratio of actual motor speed to rated motor speed, n/n_r, and the ratio of the actual motor voltage to the rated motor voltage, U/U_r, is given by the characteristic in Fig. 5, while the relation between n/n_r and the ratio of the actual moment and rated moment, $\boldsymbol{M/M_r}$, is shown in Fig. 6.

The power section of the inverter is preceded by a control stage which, among others, contains the electronics for the generation of the variable-frequency rotating field. Two basic circuits for the generation of this field are shown in Fig. 7: both are digital phase advancers.

Figure 7a provides a presettable frequency to control a ring counter. The outputs of this counter are connected to a binary-coded-decimal (BCD) decoder. The signals that are shifted by 120° with respect to one another, and which are required for driving the power semiconductors, appear at the output of the three bistables.

The circuit in Fig. 7b is rather simpler in that it does not include a BCD decoder. The AND gate obviates any unacceptable state, but has no other effect on the operation of the phase advancer.

For a rotating field frequency of 50 Hz, the clock of both designs must be 300 Hz.

Some semiconductor manufacturers provide special ICs for the generation of the rotating field, but in state-of-the-art frequency changers microprocessors are used for this purpose.

As already stated, the output voltage must rise in direct proportion to the frequency. There are a number of ways in which the output voltage may be varied: the two most important of these, pulse-amplitude modulation—PAM—and pulse-duration modulation—PDM—will be described below.

Pulse-amplitude modulation

In the PAM method of varying the output voltage, a chopper (electronic switch), following the smoothing filter, opens and closes at a rate determined by the control stage (see Fig. 8). This results in a variable direct voltage that is in direct proportion to the duty factor of the control signal.

The level of the voltage can be adjusted

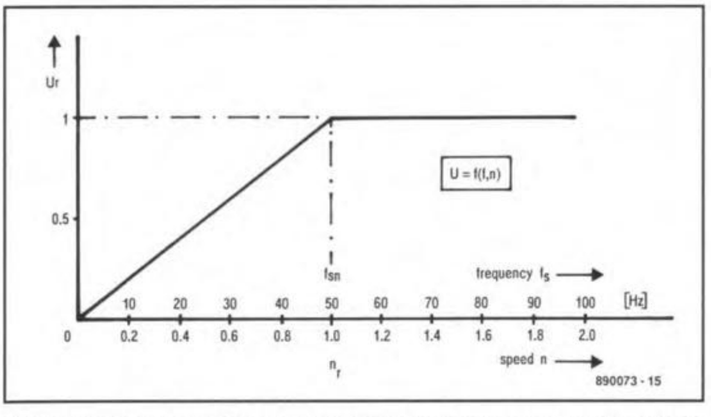


Fig. 5. Characteristic curve of the motor voltage vs rotational speed vs source frequency.

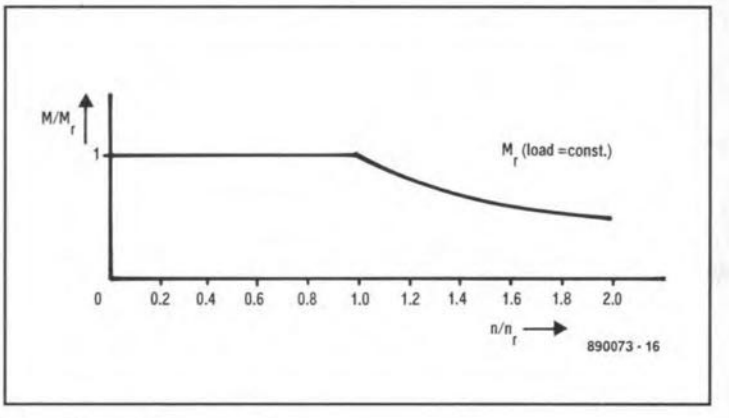


Fig. 6. Moment characteristic.

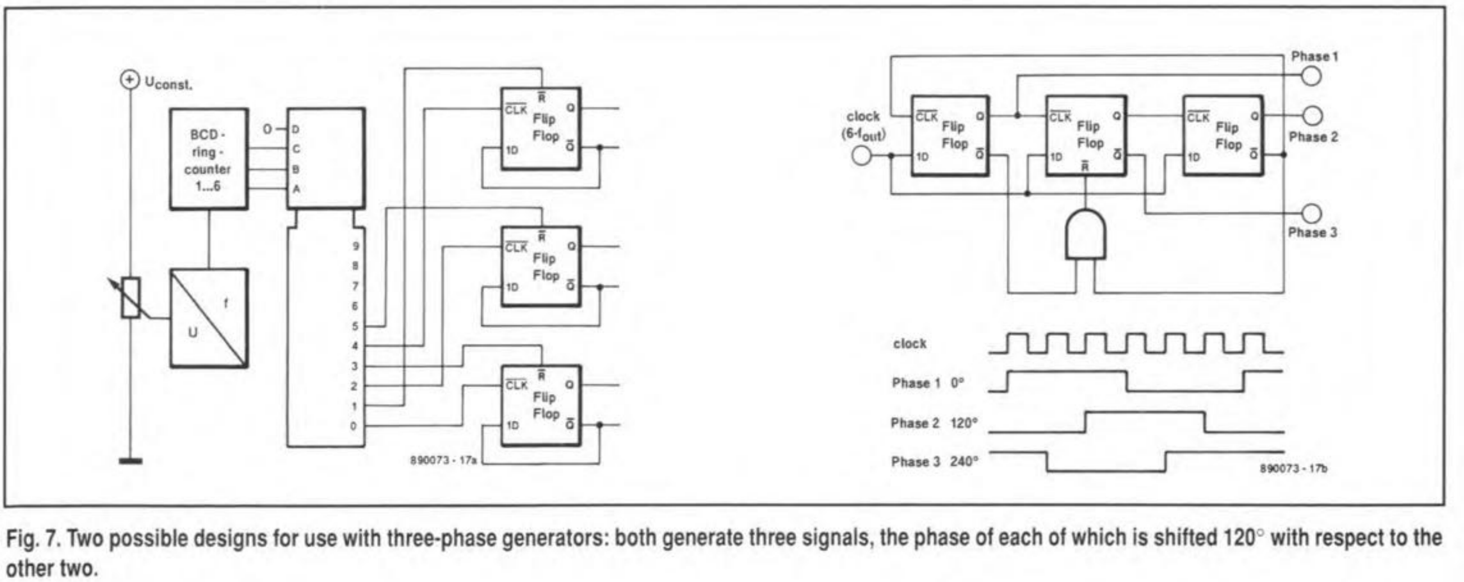


Fig. 7. Two possible designs for use with three-phase generators: both generate three signals, the phase of each of which is shifted 120° with respect to the other two.

in proportion to the frequency in a manner which ensures that the ratio of output voltage to output frequency remains constant. For a 400 V motor, this ratio must be 400 V:50 Hz = 8 V per Hz.

In the type of frequency changer described here, the output voltage is built up from six pulses per period (see Fig. 9). That voltage is not sinusoidal: it contains, apart from the fundamental frequency component, a number of harmonics of which the 5th, 7th, 11th and 13th are the most important. These harmonics cause the motor to produce spurious moments and result in additional losses in energy.

Because of the inductance of the stator, the current through the motor is rather more sinusoidal than the voltage. The waveform becomes better when the voltage is built up from a larger number of pulses per period. For instance, when there are 18 pulses per period, the output voltage and current are shaped as shown in Figure 10. The amplitude of the harmonics is then much smaller so that the motor runs rather more smoothly.

The PAM frequency changer may also be designed with a controlled rectifier instead of a chopper as shown in Fig. 11. The amplitude of the voltage across the smoothing filter can then be varied by controlling the phase angle. The main drawback of this method is the feedback on to the source (mains), which manifests itself in a poor power factor and relatively high-level harmonics. These harmonics must be filtered out, otherwise the user might run into trouble with the relevant electricity supply authority.

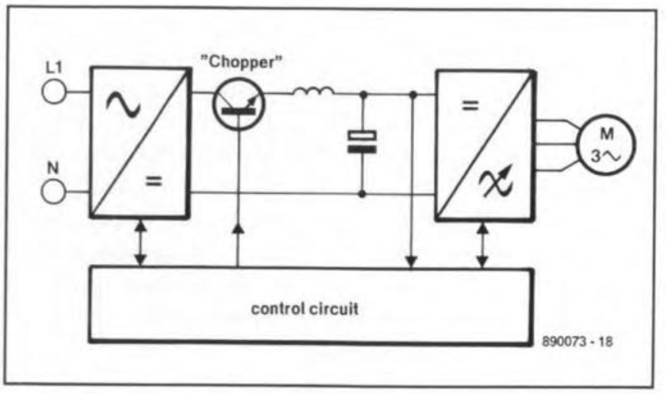


Fig. 8. Frequency changer using pulse-amplitude modulation.

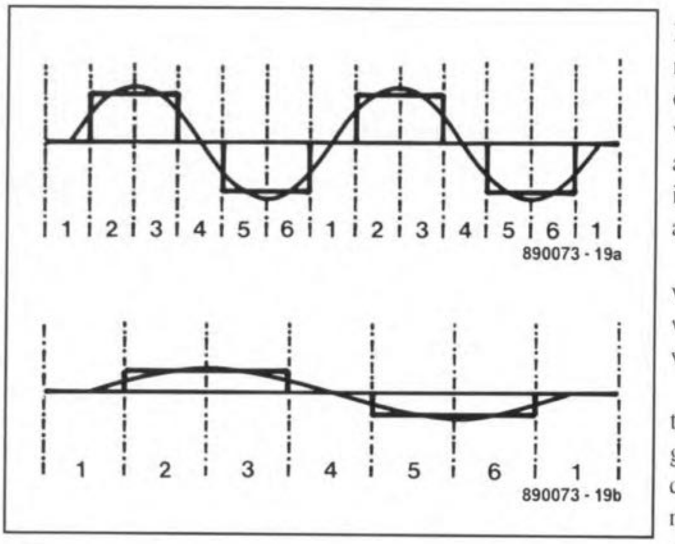


Fig. 9. Output voltage of a PAM frequency changer built up from 6 pulses per period: (a) at maximum source voltage and frequency and (b) at half the maximum source voltage and frequency.

Pulse-duration modulation

Frequency changers using pulse-duration modulation have no chopper or controlled rectifier, so that the voltage across the smoothing filter assumes a fairly constant value that is dependent only on the supply voltage and the load.

The level and shape of the output voltage are determined solely by the width of the output pulses of the inverter.

The basis of this type of modulation is a comparison between a triangular and a sinusoidal signal. The frequency of the triangular signal is much higher than that of the sine wave (see Fig. 12). The resulting pulse train is the control instrument.

As before, because of the integrating action of the stator inductance,

Fig. 10. Output voltage of a PAM frequency changer built up from 18 pulses per period (a) and corresponding current (b).

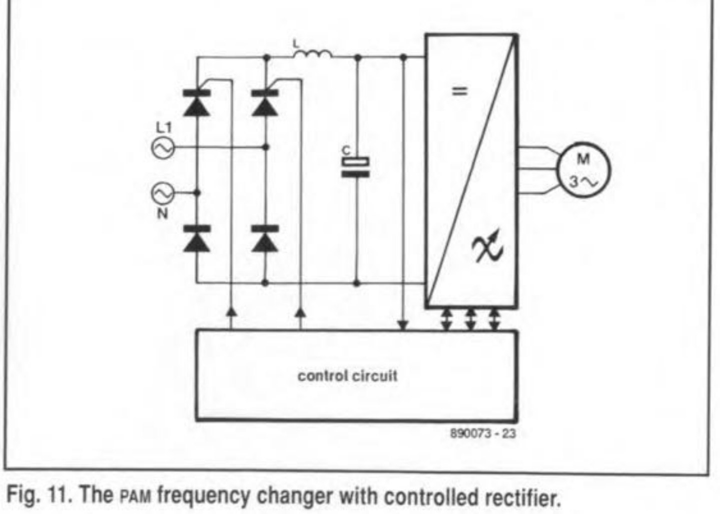


Fig. 11. The PAM frequency changer with controlled rectifier.

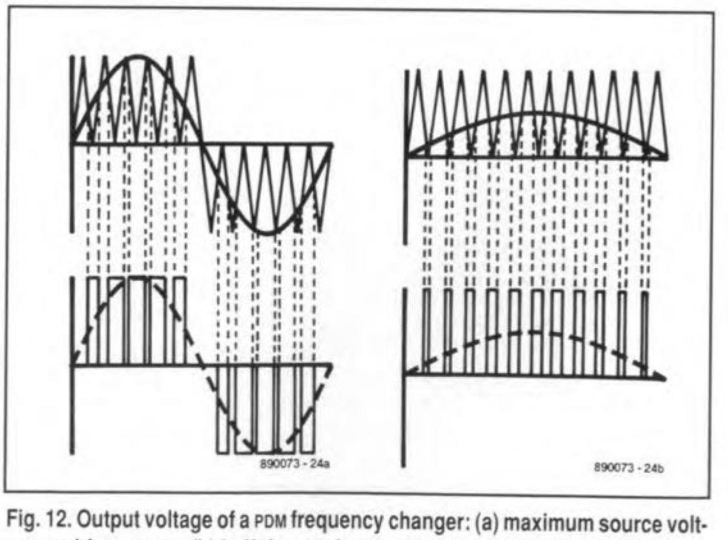


Fig. 12. Output voltage of a PDM frequency changer: (a) maximum source voltage and frequency; (b) half the maximum source voltage and frequency.

the current is sinusoidal for all practical purposes.

The design of the control circuitry is much simplified by the availability of special ICs that contain the comparator, sine wave generator and triangular-wave generator. Again, the design may also be based on the use of microprocessors.

The frequency of the sinusoidal voltage determines the output frequency of the frequency changer and thus the speed of the motor. When the frequency of the triangular signal is constant, pulses are generated that depend on the output frequency. At low output frequencies, the number of pulses per period is relatively high, which results in a near-sinusoidal current through the motor. When the output frequency is high, the waveform of the current deteriorates.

High-voltage switching transistors may be used in the power section of inverters as long as the output power does not exceed 50 kVA. Modules are available from some manufacturers that contain all six transistors required for a three-phase bridge circuit.

When the output power exceeds 50 kVA, thyristors must be used. Figure 13 shows one phase circuit of the power section of a thyristors-based inverter; the other two circuits are identical. Since thyristors can not be switched off in a simple manner, some additional components are required.

Each phase circuit needs four thyristors, of which two, Th1 and Th2, switch the current through the motor. The other two, Th3 and Th4, are required for the so-called commutation process, which is described below.

Commutation

During forward conduction all the junctions of a thyristor are forward biased. To be able to turn off the device, the charge carriers must be removed and this is usually done by applying a reverse voltage across the thyristor, a process known as commutation. The process wil be described with the aid of the circuit shown in Fig. 14, which is part of that in Fig. 13.

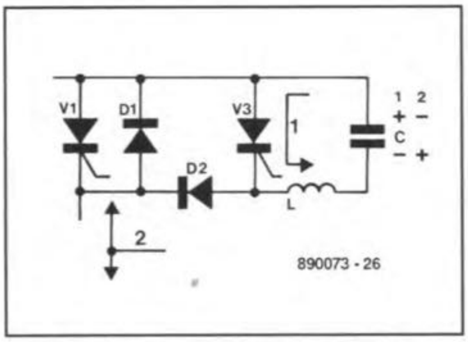


Fig. 14. Simplified commutation circuit (detail of Figure 13).

The simplified circuit in Fig. 14 consists of two thyristors, a commutation *LC* network, which forms a series resonant circuit, and two diodes.

In order to turn off Th1, Th3 must be fired. At first, capacitor C is charged as shown (1). This results in a sinusoidal current in the *LC* circuit. The first half period of that current flows through Th3, which then fires. When the current reaches a zero crossing, Th3 is turned off and the polarity of C is reversed (2). The next half period of the current flows through the diode and Th1 or the motor. For a brief instant, the motor current flows through the *LC* circuit: no current then flows through Th1 and it turns off.

The remaining components in Fig. 14 protect the power semiconductors or are needed to prepare the commutation process.

The operation of this type of inverter depends on the voltage across the motor and the inverter is, therefore, called a voltage inverter. The level of the current through the motor depends entirely on the load on the motor.

It is also possible to base the frequency changer on the processing of the current, in which case the load on the motor determines the voltage across the motor. This type of frequency changer is called a current-based frequency changer

The design of the power section of a current-based frequency changer is shown, greatly simplified, in Fig. 15. Note that the smoothing filter does not contain a capacitor. The controlled rectifier at the input determines the r.m.s. value of the current through the smoothing filter. The inductor smoothes the direct current. Before it is fed to the stators of the motor, the current through the filter is divided in the inverter in a manner to ensure that a rotating field is induced. The current through, and the voltage across, the motor are shown in Fig. 16. Note that in this type of frequency changer the current waveform is rectangular, whereas the voltage is near-sinusoidal.

Because of the controlled rectifier, the power factor at the source end of a current-based frequency changer is not constant, but varies in accordance with the load. Furthermore, this type of frequency changer can not be used in parallel operation of motors, because the commutation capacitor in the inverter must be adapted to the rating of the motor.

Since in a current-based frequency changer the motor is part of the commutation network, such a frequency changer, in contrast to a voltage-based changer, can not be operated without the motor. On the other hand, the design of the inverter for a current-

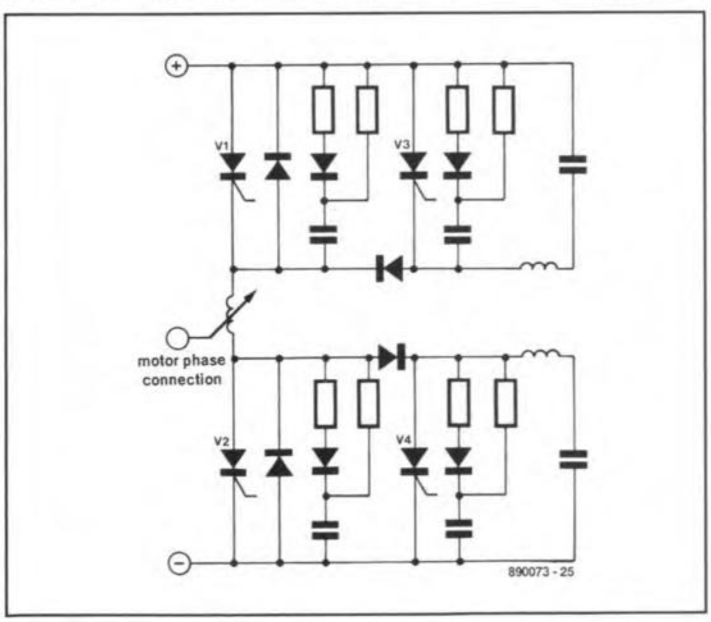


Fig. 13. Diagram of one phase circuit of a frequency changer based on thyristors, also called voltage-based frequency changer.

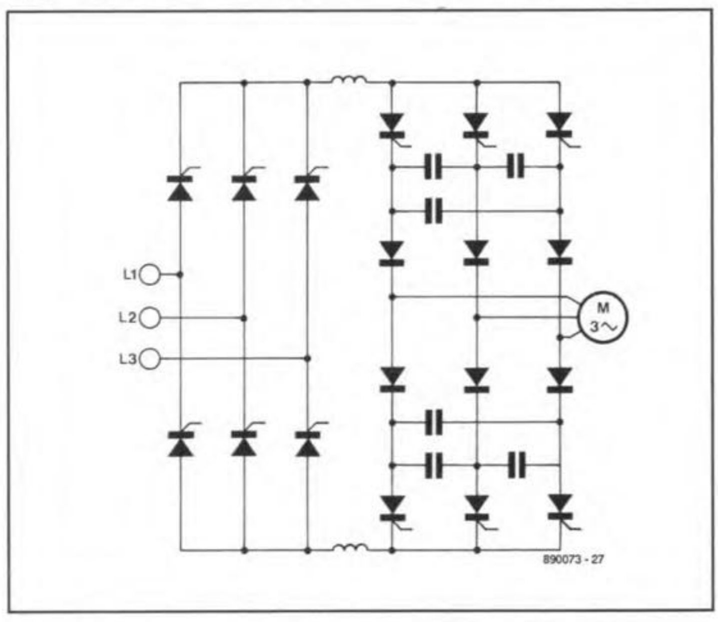


Fig. 15. Circuit diagram of the power section of a current-based frequency changer.

based frequency changer is much simpler than that for a voltage-based model. Furthermore, the commutation process is simply effected by the inverter and the rotating field of the motor without the need of additional components.

However, the great advantage of a current-based frequency changer is not its simpler design, but the possibity of feeding back the brake horsepower into the source without the need of additional components. When the motor thus operates as a generator, the current through it is reversed and fed to the controlled rectifier via the smoothing filter. The rectifier then functions as an inverter, which returns the energy to the source.

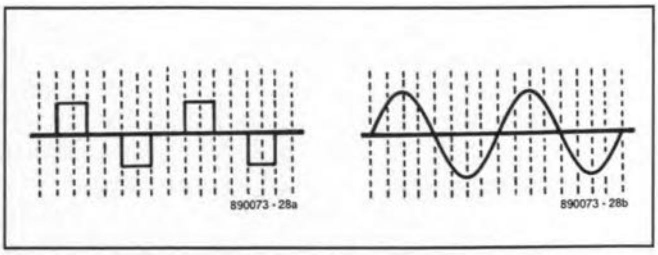


Fig. 16. Current through, and voltage across, a three-phase motor operated with the aid of a current-based frequency changer.

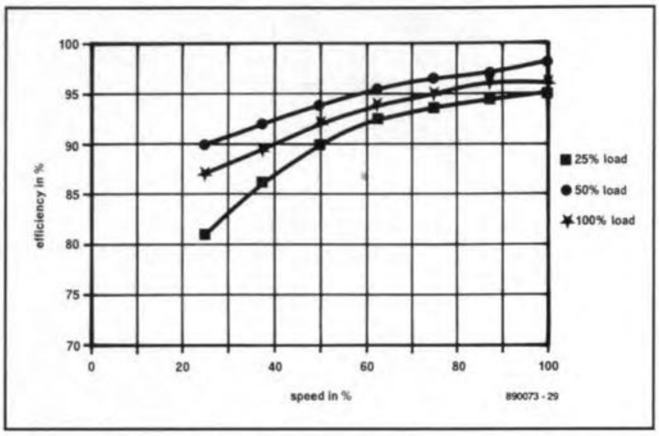


Fig. 17. Typical efficiency curves of a voltage-based frequency changer.

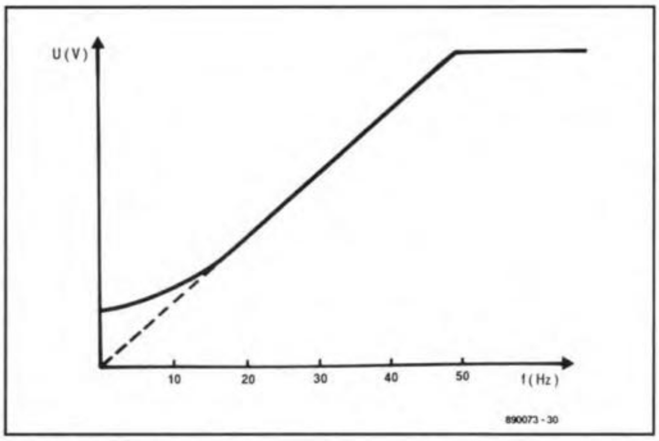


Fig. 18. Output voltage vs frequency characteristic when the voltage is raised at low frequencies.

Efficiency

The efficiency of the modern frequency changers discussed in this article is pleasantly high: depending on the design and technology employed, it lies somewhere between 93% and 97%. Typical efficiency curves of a voltage-based frequency changer operating with various loads are given in Fig. 17.

The overall dissipation includes losses in diodes, thyristors, inductors and control sections. Most of the dissipation, however, occurs in the commutation process. The overall dissipation is, therefore, highly dependent on the number of commutations per unit time. Thus, although high pulse rates reduce the harmonic content, they increase the dissipation.

At frequencies below 10 Hz, the potential drop across the resistance of the motor winding reduces the flux and the moment. In these circumstances, this may be compensated by a higher than proportional rise in the output voltage as indicated by the characteristic in Fig. 18.

Miscellaneous

When adapting your own requirements to the observations and discussions in this article, bear in mind that a number of components are shown for very large loads. In no circumstances, for instance, must the large capacitors shown in the smoothing filters be connected directly across the mains or other high-voltage source since at the moment of first switch-on they form a short circuit. Such high-value capacitors must be charged via a suitable resistor. Only when they have been charged can they be connected to the input source and even then via a suitable protection circuit.

The potential across the smoothing filter rises to the peak value of the source voltage. The capacitor(s) will retain a lethal charge even after the source has been disconnected.

The commutation process causes high voltage peaks, which should be borne in mind during test and measurement.

As a rule, all components in the power section carry a potential to earth. Only very small frequency changers can be isolated from the source (mains) by a suitable transformer. Again, this should be borne in mind during test and measurement. Also, test instruments whose 'zero potential' is connected to their case (as in a number of oscilloscopes) must be connected to the mains via an isolating transformer. There exists then, however, a potential between the case and earth. **BE WARNED** that in certain circumstances it may be dangerous to touch the case.

The voltage peaks in the inverter may reach values that exceed the isolation breakdown voltage of the test instrument. It is therefore important that for each instrument at least the maximum input voltage and the highest permissible voltage with respect to earth or to case are known and strictly observed. Furthermore, in order to judge whether a given instrument is suitable for certain measurements, its principle of operation must be known. This is particularly so when alternating voltages and currents are measured whose waveform does not match that with which the instrument has been calibrated.■

FURTHER READING:

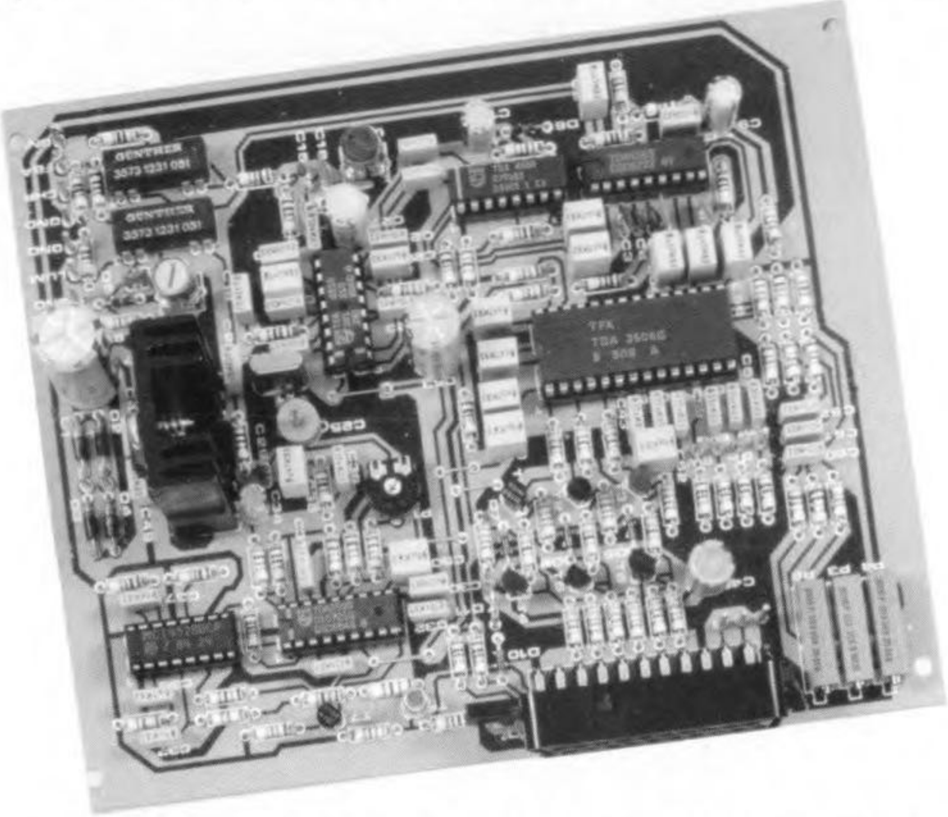
Power Electronics Handbook
by F.F. Mazda
ISBN 0 408 03004 6
Butterworth Scientific Ltd
Westbury House,
Bury Street
GUILDFORD GU2 5BH
England
(*Elektor Electronics* June 1990)

Power Electronics
by M.H. Rashid
ISBN 0 13 686619 0
Prentice-Hall
66 Wood End Lane
HEMEL HEMPSTEAD HP2 4RG
England
(*Elektor Electronics* October 1988)

Electric Machinery
by Peter F. Ryff
ISBN 0 13 248691 1
Prentice-Hall
Wood End Lane
HEMEL HEMPSTEAD HP2 4RG
England
(*Elektor Electronics* January 1989)

Solving Problems in Electrical Power and Power Electronics
by H.F.G. Gwyther
ISBN 0 582 28644 1
Longman Scientific & Technical
Longman House
Burnt Mill
HARLOW CM20 2JE
England
(*Elektor Electronics* March 1989)

S-VHS/CVBS-TO-RGB CONVERTER



PART 2: CIRCUIT DESCRIPTION AND CONSTRUCTION

Following last month's introduction into the main characteristics of the Super-VHS system, we close off the article with details of a practical converter circuit that allows an S-VHS VCR or camcorder to be connected to the RGB inputs of a colour TV or monitor. The circuit presented here forms a state-of-the-art approach to all-analogue picture standard conversion, and is based on the latest in IC technology available for this purpose.

H. Reelsen

It seems odd that the introduction of S-VHS camcorders and video cassette recorders (VCRs) last year has not been followed by more TV sets with separate chrominance and luminance inputs. After all, these recorders need a suitable display to match their advanced features. True, some buyers will opt for the fairly expensive TV receivers that have separate colour processing facilities, but many others are either not prepared to pay the current high price for such a set or net yet willing to replace their existing set. What can they do unil they have acquired a suitable new set?

The circuit presented here converts the separate colour signals supplied by an S-VHS source into the three basic colour signals, red, green and blue (RGB), which may be applied to the respective inputs of a TV set fitted with a SCART socket or separate RGB sockets. The performance of the integrated circuits used in the converter is so good that it is also worthwhile to have them convert CVBS (composite video) into RGB. A separate input is provided for this application, which also allows some types of RGB computer monitor to be used as a video display.

Circuit description

As shown in the circuit diagram, Fig. 4, the converter has three video signal inputs:

- **CVBS** (chrominance-video-blanking-synchronisation) with an input impedance of 75 Ω. This input is suitable for connecting to signal sources (VCRs, cameras, camcorders and home computers) that supply the standard CVBS signal level of about 1 V_{pp}.
- **Y** (luminance, or brightness) with an impedance and sensitivity of 75 Ω and 1 V_{pp} respectively. The Y signal is processed without a colour trap at a bandwidth of up to 7 MHz.
- **U/V** (chrominance, or colour information) with an input impedance and sensitivity of 75 Ω and 0.5 V_{pp} respectively. This input feeds the colour signal to the PAL decoder in the circuit.

Luminance processing

The function of the separate Y, U and V signals that together make up a colour video sig-

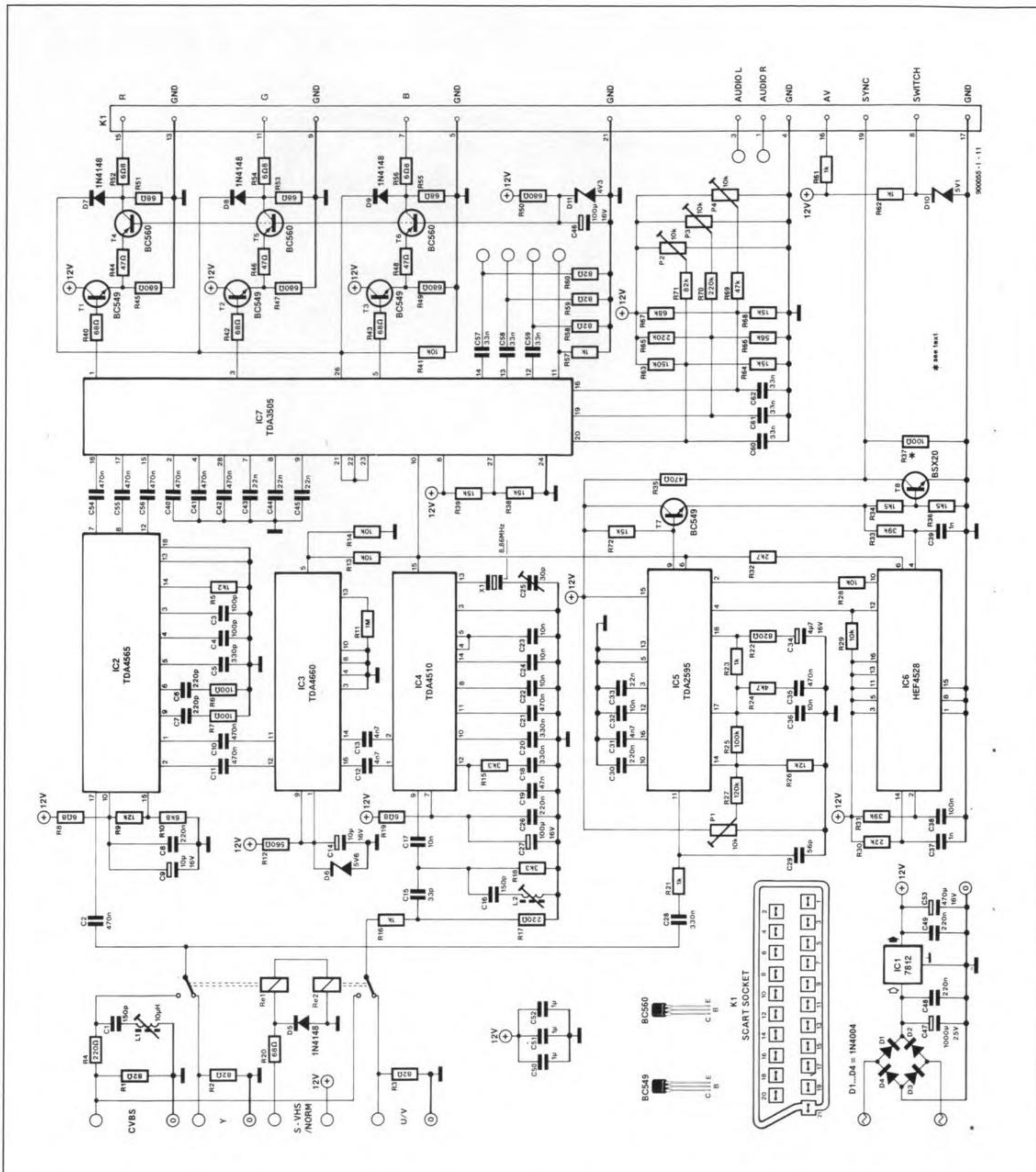


Fig. 4. Circuit diagram of the video standards converter. The unit is connected to the TV set or monitor via SCART socket K1.

nal is discussed in Part 1 of this article. The Y and U/V inputs are used with S-VHS equipment. The CVBS input may be connected to equipment that supplies a composite video signal. Two relays, Re1 and Re2, are used to switch between S-VHS and CVBS operation. The converter is switched to CVBS operation by applying +12 V to the S-VHS/NORM control input. S-VHS operation is selected by leaving the input open-circuited. Relay Re1 then feeds the Y (luminance) signal to IC2 via coupling capacitor C2. As shown in the block diagram in Fig. 5, the TDA4565 contains a colour transient improvement (CTI) circuit and a delay line for the Y signal. This delay line is an essential part in any colour TV set because the luminance signal has a much greater bandwidth than the chrominance signal and hence requires a delay of about 800 ns. A number of gyrators in the TDA4565 allow delay times between 690 ns and 960 ns to be set in steps of 90 ns with the aid of a control voltage applied to pin 15. In the present circuit, the delay is set to 780 ns by potential divider R9-R10. Coupling capacitor C56 feeds the delayed Y signal supplied by pin 12 of the TDA4565 to pin 15 of the colour matrix circuit, a TDA3505 (IC7). The delayed Y signal has an amplitude of about 0.5 V_{pp}.

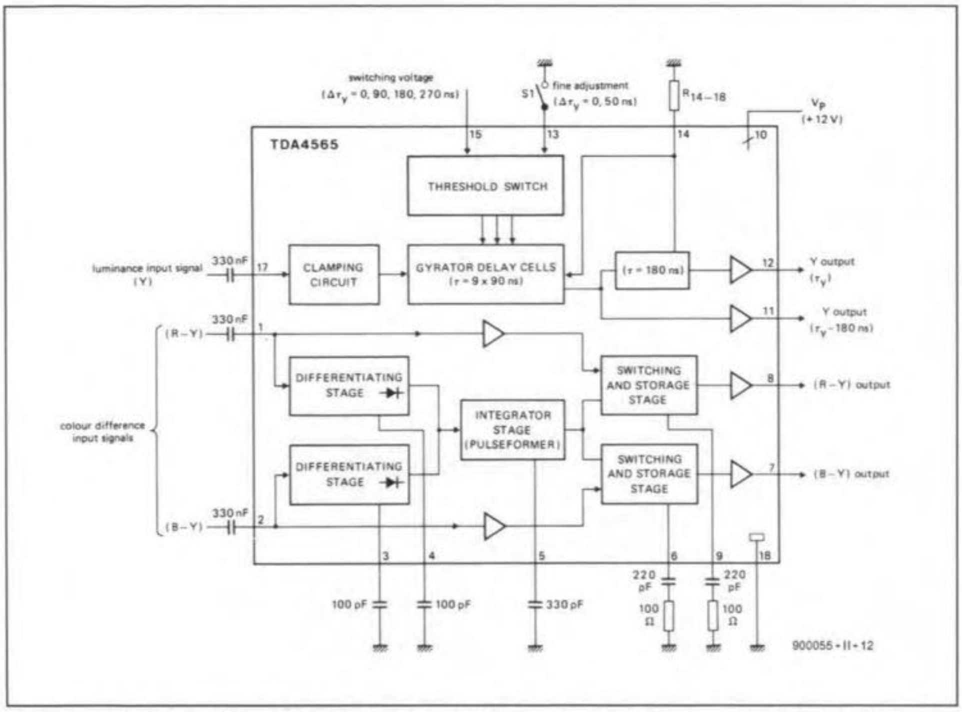


Fig. 5. Block diagram of the TDA4565 Colour Transient Improvement (CTI) circuit.

Chrominance processing

The chrominance signals are passed to the decoder via the contact of Re2. Before they arrive at the PAL decoder, a TDA4510 (IC4), the U/V signals are attenuated by R16-R17, and taken through a high-pass filter composed of C15-C16-L2.

The TDA4510 was originally designed for use with a glass delay line which serves to store the chrominance signal of the previous picture line. In the present circuit, the TDA4510 works without this crucial part whose delay time of one picture line enables the colour correction operation in the PAL TV system to correlate the colour information in two successive picture lines. In a PAL TV receiver, the R–Y and B–Y components modulated on the 4.43 MHz colour subcarrier are delayed and subsequently added to the undelayed signals. Since this addition is critical in respect of phase and amplitude, a preset and a small inductor are fitted to optimize the adjustment which, unfortunately, requires a calibrated PAL signal source.

The recently introduced TDA4660 provides a welcome alternative to the glass delay line and at the same time eliminates the associated complex phase and amplitude adjustments. The baseband delay element in the TDA4660 may be used by configuring the PAL decoder as shown in the circuit diagram. The demodulated colour difference signals at output pins 1 and 2 of IC4 are applied to the respective inputs of the CCD-based analogue shift register in the TDA4660 (see the block diagram in Fig. 6). After the shift operation, the delayed signal and the undelayed signal are added in the IC to give the conventional R–Y and B–Y components. The clock for the CCD register is provided by a PLL (phase-locked loop) circuit contained in the TDA4660. The reference clock of the PLL is formed by the line frequency, obtained from the super-sandcastle pulse applied to the chip via R13-R14. The origin and the function of the super-sandcastle pulse is discussed further on in this article.

CTI function

The colour difference signals, R–Y and B–Y, are applied to the inputs, pins 11 and 12, of the baseband delay element, IC3. The typical signal levels are 1.0 V_{PP} at the R–Y input (pin 11) and 1.3 V_{PP} at the B–Y input (pin 12). An oscilloscope connected to these IC pins will reveal sluggish rise and fall times of the colour difference signals as a result of, say, the standard colour bar test chart. This is caused mainly by the limited bandwidth (about 1 MHz) of the chrominance signal. The bandwidth is reduced even further (to about 0.5 MHz or smaller) when a normal VHS tape is played back. Obviously, this makes the signal edges even slower and results in degraded colour transient definition, or, in other words, a picture that is not very sharp. In not a few cases, the picture quality from a VCR is degraded further by moiré effects in the already blurred colour transients. As already explained in Part 1, this moiré is caused mainly by insufficient suppression of the colour subcarrier sidebands.

Fortunately, the picture quality can be improved considerably by a colour transient improvement (CTI) chip. Here, the TDA4565 (IC2) is used in a standard application circuit. The way in which CTI is implemented without introducing overshoot and additional noise is discussed below.

The TDA4565 detects a colour transient by differentiating the colour difference signals. This is achieved by an internal difference amplifier and capacitors C3 and C4. When a transient is detected, an internal pulse shaper, which uses C5 as an external part, is actuated. The pulse shaper in turn causes the input signal to be stored in a sample-and-hold circuit which retains the current signal level until the transient is over. Next, 100 ns pass before the new level is supplied. The sample-and-hold function is implemented by external components R6, R7, C6 and C7. The re-shaped colour difference signals at output pins 7 and 8 of the TDA4565 are fed to the matrix circuit via a pair of coupling capacitors, C54 and C55.

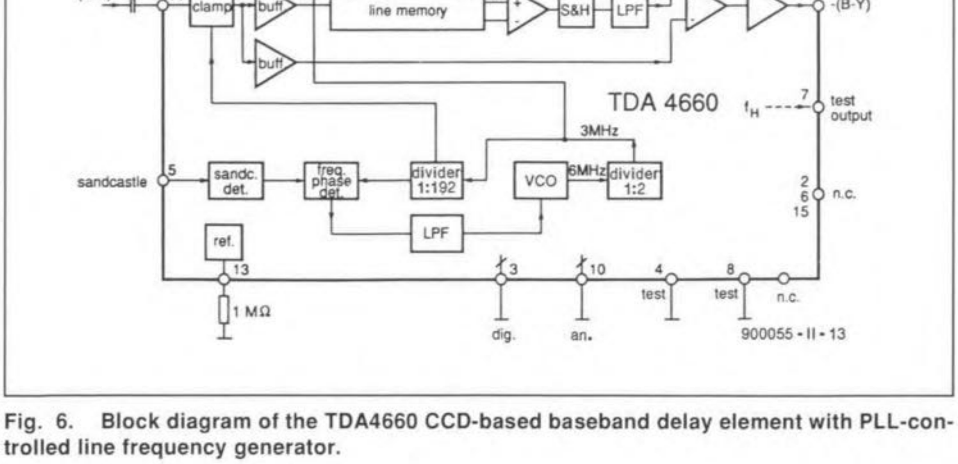


Fig. 6. Block diagram of the TDA4660 CCD-based baseband delay element with PLL-controlled line frequency generator.

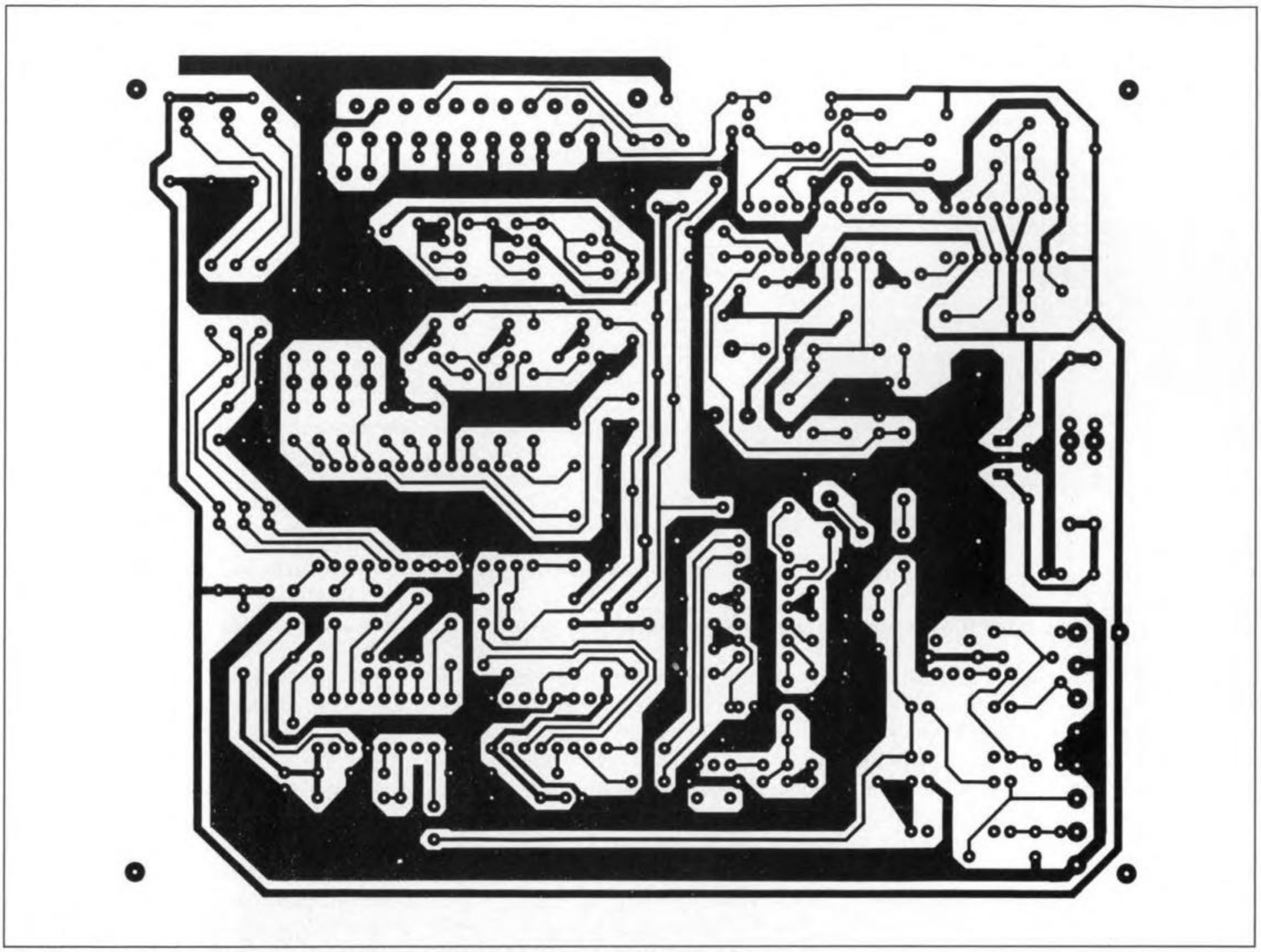


Fig. 7a. Track layout (mirror image) of the printed-circuit board for the video converter.

RGB output circuit

The colour matrix circuit is based on another video IC from Philips Components: the TDA3505 (IC7). In this, the luminance and chrominance signals meet (in S-VHS mode) or meet again (in CVBS mode). The basic colours, R, G and B, are recovered, by a summing operation, from the colour difference signals and the luminance (Y) component. The picture settings, contrast, brightness and colour saturation, are adjusted by direct voltages that determine the bias and the gain at a number of points in the matrix. Here, the relevant components are R63-R71 and presets P2, P3 and P4. The presets are used to adjust the brightness (P2), the contrast (P3) and the colour saturation (P4). The multiturn presets on the circuit board may, of course, be replaced by front-panel mounted potentiometers to give a continuous control range rather than fixed settings.

Two-stage level shifters/buffers are required at the outputs of the matrix because these do not supply levels down to 0 V, and are not capable of driving a 75-Ω load direct. The buffering and level shifting are achieved with three combinations of an emitter follower and a common-base amplifier (T1-T6). The output impedance of the three drivers is 75 Ω.

Each colour output driver has a diode which allows the operating point of the two-transistor stage to be monitored via pin 26 of the TDA3505. The operating point is monitored and, if necessary, corrected, during the vertical blanking interval, i.e., when the scanning beam in the TV set is quenched. The direct voltage required for this function is stored in capacitors C40, C41 and C42 during the current picture. The matrix circuit

COMPONENTS LIST

Resistors:

6	82Ω	R1;R2;R3;R58;R59;R60
2	220Ω	R4;R17
1	1k2	R5
3	100Ω	R6;R7;R37
5	6Ω8	R8;R19;R52;R54;R56
2	12kΩ	R9;R26
1	6k8	R10
1	1MΩ	R11
1	560Ω	R12
5	10kΩ	R13;R14;R28;R29;R41
2	3k3	R15;R18
6	1kΩ	R16;R21;R23;R57;R61;R62
7	68Ω	R20;R40;R42;R43;R51;R53;R55
1	820Ω	R22
1	4k7	R24
1	100kΩ	R25
1	120kΩ	R27
1	22kΩ	R30
2	39kΩ	R31;R33
1	2k7	R32
2	1k5	R34;R36
1	470Ω	R35
5	15kΩ	R38;R39;R64;R68;R72
3	47Ω	R44;R46;R48
4	680Ω	R45;R47;R49;R50
1	150kΩ	R63
2	220kΩ	R65;R70
1	56kΩ	R66
1	68kΩ	R67
1	47kΩ	R69
1	82kΩ	R71
1	10kΩ preset H	P1
3	10kΩ multiturn preset	P2;P3;P4

(continued →)

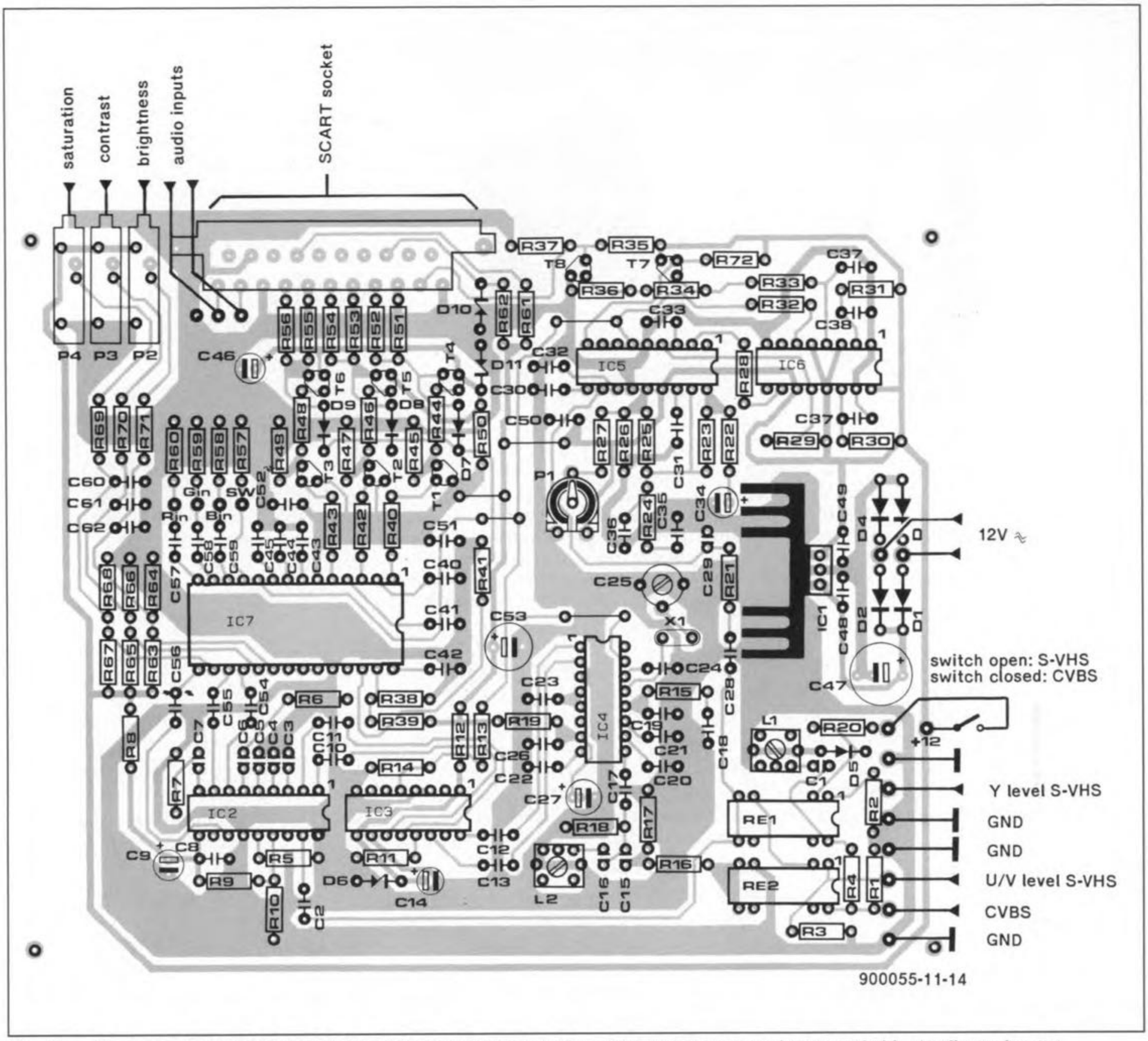


Fig. 7b. Component mounting plan of the printed-circuit board, and connections to external parts and video/audio equipment.

COMPONENTS LIST

Capacitors:

2	150pF	C1;C16
11	470nF	C2;C10;C11;C21; C35;C40;C41; C42;C54;C55;C56
2	100pF	C3;C4
1	330pF	C5
2	220pF	C6;C7
5	220nF	C8;C26;C30;C48; C49
2	10µF 16V radial	C9;C14
3	4n7	C12;C13;C31
1	33pF	C15
6	10nF	C17;C22;C23; C24;C32;C36
3	330nF	C18;C20;C28
1	47nF	C19
1	30pF trimmer	C25
2	100µF 16V radial	C27;C46
1	56pF	C29
4	22nF	C33;C43;C44;C45
1	4µ7 16V radial	C34
2	1nF	C37;C39
1	100nF	C38
1	1,000µF 25V radial	C47
3	1µF	C50;C51;C52
1	470µF 16V radial	C53
6	33nF	C57–C60

Semiconductors:

4	1N4004	D1–D4
4	1N4148	D5;D7;D8;D9
1	5V6 0.4W zener diode	D6
1	5V1 0.4W zener diode	D10
1	4V3 0.4W zener diode	D11
4	BC549B	T1;T2;T3;T7
3	BC560C	T4;T5;T6
1	BSX20	T8
1	7812	IC1
1	TDA4565	IC2
1	TDA4660	IC3
1	TDA4510	IC4
1	TDA2595	IC5
1	HEF4528	IC6
1	TDA3505	IC7

Inductors:

1	10µH adjustable; Toko 119 LN-A3753	L1
1	50µH adjustable; Toko 119 LN-A5783	L2

Miscellaneous:

2	12-V SPDT DIL reed relay	Re1;Re2
1	PCB-mount SCART socket	K1
1	quartz crystal 8.867238 MHz (HC18/U)	X1
1	heat-sink for IC1	
22	solder pins	
1	printed-circuit board	900055

Fig. 8. S-VHS equipment is gaining rapid acceptance. Pictured to the right is JVC's Super-VHS compact recorder with built-in LCD screen. The recorder is claimed to be the world's lightest and smallest at a weight of only 530 g and a size of 131×58×118 mm. The associated Super-VHS camera with stereo sound has a size of 39×69×122 mm. The camera and the recorder are part of the SC-F007 mini-video system, kindly put at our disposal by JVC Holland.

recognizes the vertical blanking period with the aid of the super-sandcastle pulse.

The SCART socket that supplies the RGB output signals also carries the (stereo) sound signals via pins 3, 1 and 4, and the AV and SWITCH voltages (+12 V and +5 V for automatic switch-over to AV and RGB mode respectively).

CVBS mode

The operation of the circuit in CVBS (composite video) mode is much simpler than in S-VHS mode. When the S-VHS/NORM control input is connected to +12 V, both relays are actuated. Like the chrominance signal, the CVBS signal is applied direct to the colour filter, so that the PAL decoder receives the colour components, which, obviously, the Y channel must not be allowed to 'see'. The filtered composite signal is applied to IC2 after passing a colour trap composed of R4 and tuned circuit L1-C1. The CVBS (or Y) signal is 'tapped' behind Re1 and fed to the synchronization separator, IC5, via a low-pass filter, R21-C29.

Power supply

The 12-V power supply on the board is conventionally based on a rectifier, D1-D4, a smoothing capacitor, C17, and a voltage regulator, IC1. The input of the supply may be provided with an alternating voltage between 10 V and 12 V.

Syncs and sandcastles

The horizontal sync generator and sync separator is formed by IC5, a TDA2595. This IC also generates the previously mentioned sandcastle pulse.

When pin 9 of the TDA2595 is connected to +12 V via a 15-kΩ resistor, the complete synchronization signal is available as positive-going pulses with a swing of 12 Vpp. Inverter T8 is driven by T7, an emitter follower. The open-circuited signal level at the SYNC output of the SCART socket is set to about 2 Vpp by voltage divider R35–R37 at the collector of T8. When this output is loaded, the signal level drops to about 1 Vpp. When a multi-sync monitor with a TTL-compatible sync input is used, resistor R37 must be changed to 390 Ω.

The TDA2595 requires the horizontal and the vertical blanking pulse to generate the super-sandcastle pulse. This four-level pulse contains the following timing information:

- 0 V = picture period and reference level
- +2.5 V = vertical blanking
- +4.5 V = horizontal blanking
- +11 V = burst gate

The burst gate is obtained from the PLL-controlled line frequency generator in the TDA4660. It enables the PAL decoder to time the insertion of the 4.43 MHz colour burst in the horizontal blanking period. Since the horizontal and vertical blanking pulses are normally generated in the deflection circuits of the TV set, they must be generated separately in the converter. This is achieved by a dual monostable, IC6.

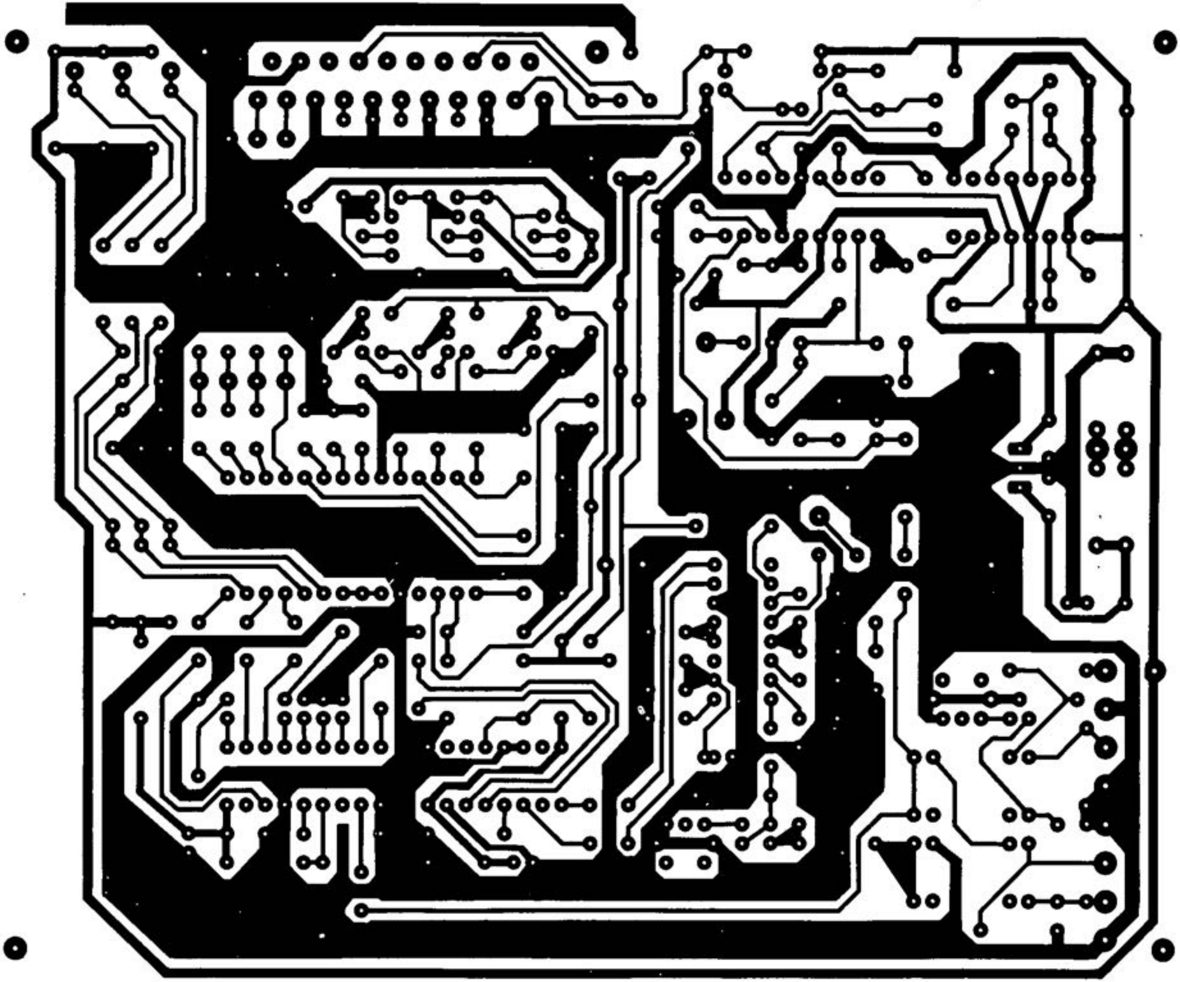
The positive-going composite synchronization signal at the emitter of T7 is passed through low-pass filter R33-C39, so that the vertical synchronization component remains. It triggers one of the monostables via pin 4. At the output, a 1.2-ms long pulse appears, which is mixed with the sandcastle pulse via R32. The output signal of the horizontal sync oscillator (at pin 4 of IC5) is fed to the second monostable in IC6. This supplies a 10-µs long pulse which is fed back to the TDA2595 for use as the horizontal blanking level in the sandcastle pulse.

Construction and adjustment

Although the circuit is relatively complex, its construction on the single-sided printed-circuit board shown in Fig. 7 is straightforward. Start the construction by fitting the five wire links on the board. The voltage regulator, IC1, must be bolted to a fairly large, vertically mounted heat-sink before its terminals are soldered. If the SCART socket has mounting holes in the flanges, they must be used to secure the plastic body to the printed-circuit board with the aid of two small screws (M3) and nuts. Some SCART sockets have snap-in arms at the sides for which holes must be drilled in the PCB. Do not forget to set the three multiturn presets to the centre of their travel before or after they are mounted: else, strange picture effects may occur when the converter is first switched on, and you may have a hard time finding the cause of the problem, when there is nothing wrong with the circuit.

On completion of the solder work, inspect the printed-circuit board very carefully. Check the orientation of all ICs, diodes and electrolytic capacitors against the overlay printed on the board and shown in Fig. 7b.

Apply power to the converter and check that its current consumption is about 350 mA at 12 V. Next, adjust preset P1 until the PLL runs free at the line frequency, 15,625 Hz (64 µs), which can be measured at pin 4 of IC5. Apply a colour input signal, and adjust trimmer capacitor C25 until the monitor switches to colour. In most cases, the colour will be on already with the trimmer set to roughly half-way of its travel. Check that the 8.86 MHz oscillator starts properly by switching the converter on and off a few times. The colour should come on immediately after switching on. If it does not, carefully re-adjust the trimmer. Finally, adjust the colour trap, L1. Apply a CVBS signal and adjust the inductor for minimum chrominance subcarrier amplitude. This measurement is best carried out with an oscilloscope connected to pin 12 of IC2. When an oscilloscope is not available, adjust L1 for minimum moiré interference in the colour picture. ■



400-watt laboratory power supply

October 1989 and November 1990

A number of constructors of this popular project have brought the following problems to our attention.

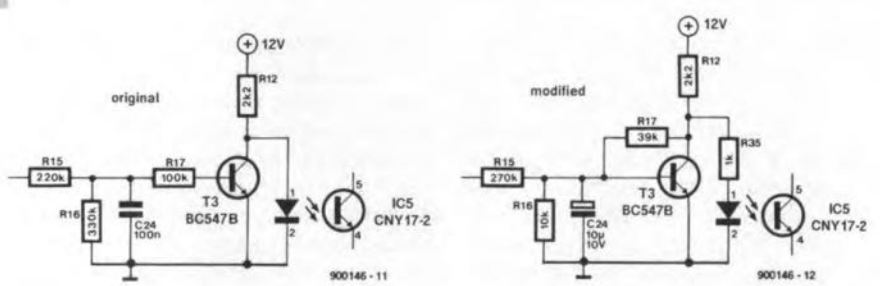
1. The onset point of the current limit circuit lies at about 3 A, which is too low. Solve this problem by replacing T1 with a Type BC517 darlington transistor, and R20 with a 82kΩ resistor.

2. Depending on the current transfer ratio of the optocoupler used, the transformer produces ticking noises. This effect, which is caused by overshoot in the pre-regulation circuit, may be traced with the aid of an oscilloscope monitoring the voltage across C26 at a moderate load current. The capacitor must be charged at each cycle of the mains frequency, and not once every five cycles. The problem is best solved by reducing the amplification of the regulation circuit. Replace R17 with a 39 kΩ resistor, and create feedback by fitting it between the base and the collector of T3. Also add a resistor in series with the optocoupler. These two changes are illustrated in Figs. 1 and 2. Lower R16 to 10 kΩ, increase C24 to 10 µF, and increase R15 to 270 kΩ.

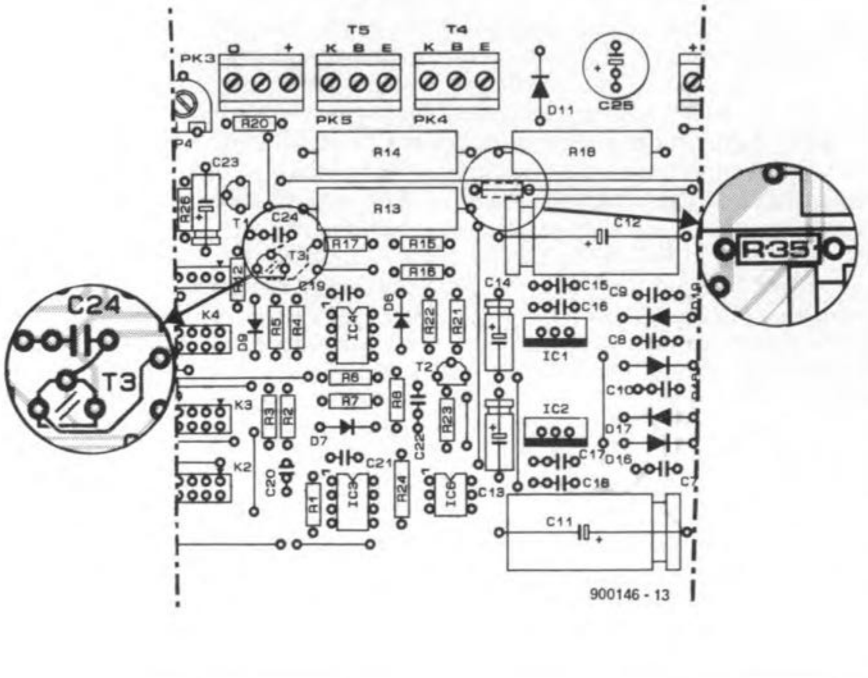
3. Excessive heating of the transformer is caused by a d.c. component in the primary winding. This is simple to remedy by fitting a capacitor of any value between 47 nF and 470 nF, and a voltage rating of 630 V, across the primary connections. This capacitor is conveniently mounted on to the PCB terminal block that connects the transformer to the mains.

4. One final point: when using LED DVMs for the voltage/current indication, their ground line must be connected to the **positive** terminal of C12.

1



2



Hard disk monitor

December 1989

In some cases, the circuit will not reset properly because the CLEAR input of IC3A is erroneously connected to ground. Cut the ground track to pin 3 of IC3, and use a short wire to connect pin 3 to pin 16 (+5 V).

Microprocessor-controlled telephone exchange

October 1990

In some cases, the timing of the signals applied to IC17 causes a latch-up in the circuit, so that the exchange does not detect the state of the connected telephones properly. Solve this problem by cutting the track to pin 1 of IC17, and connecting pin 1 to ground (a suitable point is the lower terminal of C6).

The text on the fitting of wires on the BASIC computer board (page 19, towards the bottom of the right-hand column) should be modified to read: 'Finally, connect pin 6 of K2 to pin 7 of IC3 ($\overline{Y7}$ signal).'

S-VHS/CVBS-to-RGB converter (2)

October 1990

The capacitor marked 'C37', next to R21 on the component overlay (Fig. 7b and ready-made printed circuit board), should be marked 'C39'.

In case they are difficult to obtain locally, inductors type 119-LN-A3753 (L1) and 119-LN-A5783 (L2) may be replaced with the respective types 119-ANA-5874HM and 119-ANA-5871HM, also from Toko, Inc. Suggested suppliers are Cirkit Distribution Ltd., and C-I Electronics.

EPROM simulator

December 1989

Counters IC3 and IC4 may not function properly owing to a too low supply voltage. This problem may be solved by replacing IC12 with a 7806. Alternatively, use BAT85 diodes in positions D1 and D2.

Programmer for the 8751

November 1990

The ready-programmed 8751 for this project is available at £35.25 (plus VAT) under order number **ESS 7061**, not under order number ESS 5951 as stated on the Readers Services pages in the November and December 1990 issues.

CORRECTIONS

6-metre band converter

April 1991, p. 38-43

The components list and the inductor overview in the top left hand corner of the circuit diagram should be corrected to read:

L_1, L_2 = 301KN0800.

Capacitor C_{16} (4.7 pF) must not be fitted on the board.

Finally, a few constructional tips:

- Fit a 10 nF ceramic decoupling capacitor at junction L_7-R_{36}.
- Fit a 18 kΩ resistor between the base of T_3 and ground. This reduces the Q factor of L_2, and prevents too high signal levels at the base of T_3.
- For improved tuning, inductor L_9 may be replaced by a Toko Type 113KN2K1026HM.

Multifunction measurement card for PCs

January and February 1991

We understand that the 79L08 (IC_{17}) is no longer manufactured and, therefore, difficult to obtain. Here, the IC may be replaced by a 7908, which, although physically larger than the 79L08, is pin-compatible, and should fit on the PCB.

Dimmer for halogen lights

April 1991, p. 54-58

In the circuit diagram of the transmitter, Fig. 2, pin 14 of the MV500 should be shown connected to pin 13, not to junction R1-R_2-C_2. The relevant printed-circuit board (Fig. 6) is all right.

RDS decoder

February 1991, p. 59

Line A0 between the 80C32 control board and the LC display is not used to reset the display, but to select between registers and data.

We understand that the SAF7579T and the associated 4.332 MHz quartz crystal are difficult to obtain through Philips Components distributors. These parts are available from C-I Electronics, P.O. Box 22089, 6360 AB Nuth, Holland. For prices and ordering information see C-I's advertisement on page 6 of the May 1991 issue.

S-VHS-to-RGB converter

October 1990, p. 35-40

Relays Re_1 and Re_2 must be types with a coil voltage of 5 V, not 12 V as indicated in the components list. Constructors who have already used 12-V relays may connect the coils in parallel rather than in series.

Suitable 5-V relays for this project are the 3573-1231.051 from Günther, and the V23100-V4305-C000 from Siemens.

The components list should me modified to read:

6 33nF C57-C62

MEASUREMENT TECHNIQUES (1)

By F.P. Zantis

There are many electronics/electrical practitioners who have implicit faith in their measuring instruments and believe anything displayed on the these, particularly if the display is digital. There is more to measuring electrical quantities, however, such as knowledge of the instrument being used and the purpose of the measurement. With those in mind, it is possible to select the most suitable instrument for a given measurement. Judging and interpreting the measurements are the next important steps. Basically, each and every measuring instrument produces errors. The magnitude of such errors and how to arrive at meaningful results is the object of this new series of articles.

In all measurements a number of errors may occur that can, however, be avoided or at least minimized by a proper knowledge of them. Such errors fall under four headings.

1. **Systematic errors.**

Systematic errors are in direct proportion to the quality of the instrument and are the result of inaccuracies in the design, construction, final assembly and calibration of the instrument. The more care the manufacturer has taken in these areas, the smaller the errors will be (and the higher the price of the instrument).

2. **Environmental errors**.

Environmental errors are caused by geographical position, temperature, humidity, and electric and magnetic fields. One of the most common is the electro-magnetic field around a mains transformer (see Fig. 1),

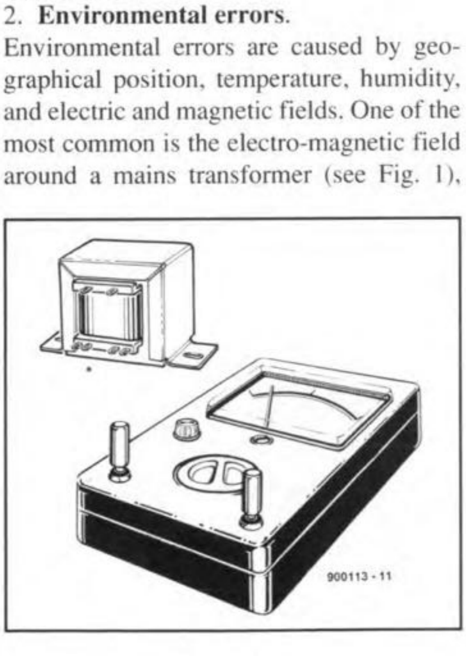


Fig. 1.

which may make a measurement at best meaningless and at worst impossible. Even the earth's magnetic field can distort a measurement: the electron beam in an oscilloscope, for instance, is deflected by it. The degree of deflection is determined by the position of the oscilloscope in the laboratory or workshop and is evidenced by the horizontal trace not being exactly parallel with the horizontal graticule lines. Screening the oscilloscope does not help and a good-quality oscilloscope is therefore provided with a trace rotation control.

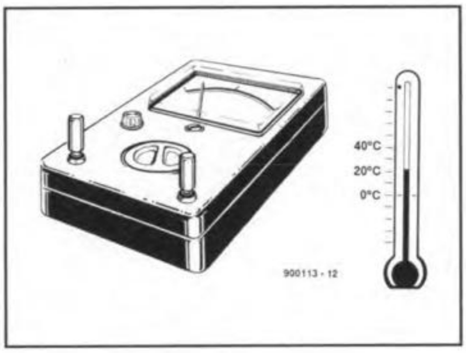


Fig. 2.

The effect of ambient temperature should also not be underestimated (see Fig. 2.). Test equipment is normally calibrated at +20 °C. When the ambient temperature differs from this by more than ±10 °C, temperature-dependent errors are likely to occur.

With analogue instruments, it may also be important whether it is used upright or lying down (see Fig. 3).

3. **Human errors.**

Human errors occur when the user is not quite clear about the purpose of the measurement or does not know enough about the test equipment he is using, or both. Typical are not knowing the input resistance when voltage measurements are carried out (error caused by incorrect loading of the voltage source) and incorrect setting of the instrument. Carelessness in taking or recording readings is another cause of errors.

4. **Applicational errors**.

Applicational errors occur when the equipment used is not suitable for the measurement to be carried out. For instance, the use of an instrument whose internal resistance is comparable to the input resistance of the circuit under test.

Also, reading errors are very common, particularly when analogue instruments are used. Such instruments should always be read with the eye at right angles to the pointer to avoid parallax errors. Some instruments have a mirror directly below the meter scale to ensure that the pointer is read correctly (it should not be seen in the mirror).

Errors in practice

Most measuring instruments are still based on analogue techniques and have analogue displays. This is so for very good reasons as

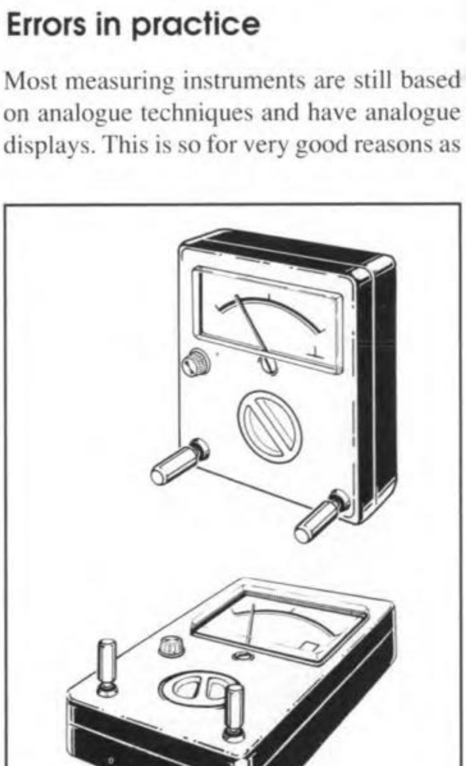


Fig. 3.

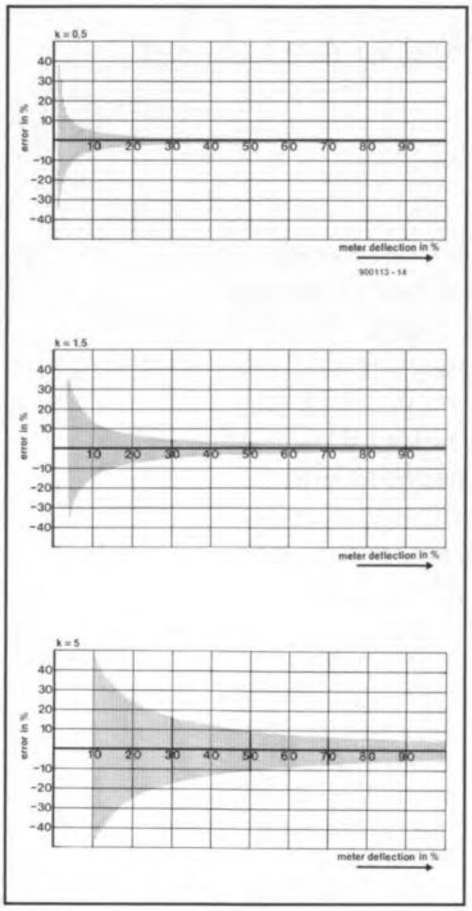


Fig. 4.

will be seen later in the series.

The accuracy of analogue instruments is normally given as a percentage (absolute error) of full-scale deflection (FSD). Note that the a.c. and d.c. ranges of multimeters usually have different accuracies. This means, for instance, that the pointer of an instrument with 1.5% accuracy and set to its 10-V range may at any given reading be off the true value by as much as 150 mV.

If, therefore, a voltage is measured as 1 V, the true value may lie between 0.85 V and 1.15 V. That represents a maximum relative error of 15%! If, in the same range, a voltage of 9 V is measured, its true value may lie between 8.85 V and 9.15 V. The maximum relative error is then 0.15/ 9×100% = 1.67%.

From these two examples it is seen that at the lower end of the range measurements are guesswork, in the middle of the range they are merely unreliable, and only at the upper end of the range does the relative error compare with the stated accuracy. This means that for greater accuracy the proper range (that is, the one where the greatest deflection is obtained) should be selected. Figure 4 shows the relative error in per cent for three instruments that have absolute errors of 0.5%, 1.5% and 5%, respectively, of full-scale deflection. In all cases, note the rapid rise of the relative error for small pointer deflections.

Matching measurand and measurement range

The importance of selecting the correct measurement range will be shown by the following example. Assume that an alternating voltage of about 20 V is to be measured and that two measuring instruments are available. Of these, one has an absolute accuracy of 0.5% and FSD ranges of 1 V, 10 V and 100 V, while the other has an absolute accuracy of 1% and FSD ranges of 3 V, 30 V and 300 V.

The first instrument must be set to the 100 V range and the relative error will be 0.5/20×100% = 2.5%. The second instrument is set to its 30 V range, where the relative error wil be 0.3/20×100% = 1.5%. It is thus seen that the less accurate instrument in this case gives the more accurate result! This is an important point to bear in mind when buying an instrument: the ranges available should be compatible with the expected measurands. When you are normally engaged in power electronics, you need different ranges from those required in radio/TV engineering.

Digital instruments

In digital instruments, the accuracy is normally related to displayed value. Furthermore, because of the analogue-to-digital conversion, the accuracy of the final digit must be added. Normally, the accuracy of the displayed value is ±1 digit, although there are instruments with an accuracy of ±2 digits and more.

The overall accuracy of a digital multimeter is typically

- for direct voltage measurements: 0.1% of measurand ±1 digit;
- for alternating voltage measurements: 2% of measurand ±7 digits;
- for direct current measurements: 0.35% of measurand ±1 digit;
- for alternating current measurements: 0.9% of measurand ±3 digits.

If the measurement of the alternating voltage of around 20 V in the earlier examples had been carried out by this digital instrument, the readings would have been

20×1.02 + 7×0.1 = 21.1 V

and

20×1.02 – 7×0.1 = 18.9 V

respectively. The relative error would have been

1.1/20×100% = 5.5%.

This shows that digital instruments are not necessarily more accurate than analogue ones as quite a few people believe. Note that the graphs in Fig. 4 apply to digital instruments as well.

Measurand and instrument

The measurand may have any one of a multitude of waveforms. Because of that, different instruments, or different methods of measuring, will give different results when such quantities are measured. Only when the basis of the measurement is known can the reading have any significance. For instance, a moving coil meter without rectifier will show the arithmetic mean of a quantity. However, the artihmetic mean of a pure alternating voltage (or current) is 0, so that there is no deflection of the pointer (see Fig. 5). If that meter is used to measure a com-

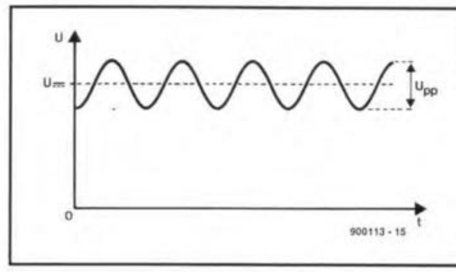


Fig. 5.

posite voltage (or current), it will only show the value of the d.c. component Care should be taken, however, because if the level of the superimposed alternating voltage (or current) is high, the meter may be overloaded in spite of that quantity not affecting the reading.

To measure an alternating voltage (or current) with a moving coil meter, a rectifier should be connected in series with the meter. This is the case, for instance, when an analogue multimeter is set to one of its a.c. ranges. The instrument will then display the mean d.c. value of the rectified quantity (see Fig. 6). The arithmetic mean of a rectified sinusoidal voltage, $U_a = 0.318U_{rms}$, and this

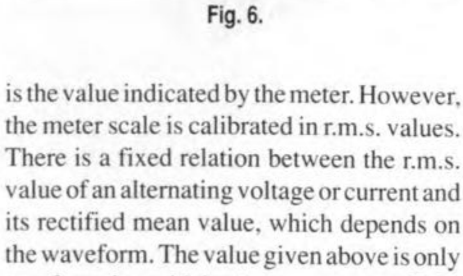


Fig. 6.

is the value indicated by the meter. However, the meter scale is calibrated in r.m.s. values. There is a fixed relation between the r.m.s. value of an alternating voltage or current and its rectified mean value, which depends on the waveform. The value given above is only true for a sinusoidal voltage or current. If the voltage or current is rectangular, the meter reading must be multiplied by 0.89 to find the r.m.s. value. When a triangular quantity is measured, the error is negligible: the true value is then only 0.36% larger than that displayed.

It might be expected that all this is true for digital instruments as well, but unfortunately that is not always so. Depending on the method of test, deviations may occur that can not really be foreseen. Some (relatively expensive) of these instruments can display the true r.m.s. value of a voltage or current. These instruments compute the r.m.s. value independently of the waveform.

Another factor that should be borne in mind is that, apart from the waveform, the frequency of a voltage or current plays a role. Most digital multimeters measure alternating quanties correctly only if the frequency is not higher than 400 Hz. Analogue instruments of about the same price perform rather better: right into the kHz range.

The r.m.s. measurement in many standard d.c. instruments is made possible by a variety of special circuits. These compute the r.m.s. value from the applied alternating voltage and convert it into a direct voltage, which is read in the usual way. These circuits are normally based on special ICs: a typical one is shown in Fig. 7 (see also Ref. 1 and Ref. 2).

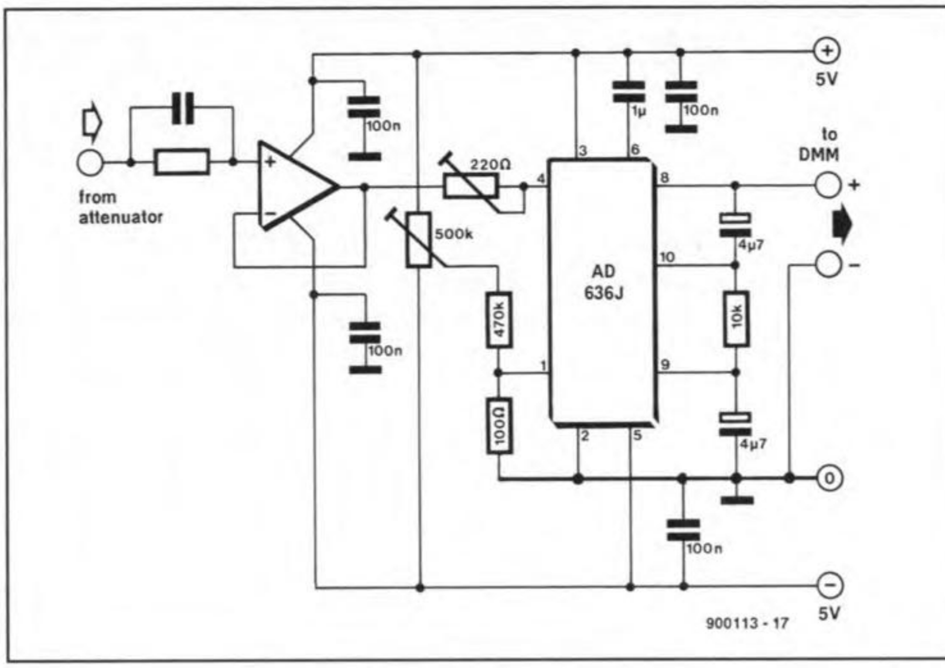


Fig. 7.

When measurements are carried out with an oscilloscope, the instantaneous value of the measurand can be seen at a glance, so that at least the peak value of a voltage or current can be determined quickly. Determining the r.m.s. value is, however, possible only if the waveform is regular and the appropriate conversion factor is known. ■

Further reading:

1. "RMS-to-DC converter", *Elektor Electronics*, July 1986, p. 38.

2. "True RMS DVM", *Elektor Electronics*, December 1989, p. 40.

3. *Electronic instruments and measurement techniques*" by F.F. Mazda, ISBN 0 521 26873 7, Cambridge University Press (*Elektor Electronics*, November 1987)

4. *Modern Electronic Test Equipment*, by Keith Brindley, ISBN 0 434 90567 4, Heinemann-Newnes (*Elektor Electronics*, November 1986).

FOUR-CHANNEL DIGITAL DELAY / PULSE GENERATOR

The Precision Pulse Generator and Digital Delay Generator recently announced by Fieldtech offers high accuracy and precision, a wide range and low jitter.

As a digital delay generator, the four outputs of the unit may be programmed for any interval between 0 s and 1000 s with 5 ps resolution. The standard timebase provides 25 ppm accuracy, but 1 ppm is available as an optional extra. The jitter of any output is less than 50 ps plus 1 part in 10^8 of the programmed delay. All outputs return to their pre-trigger levels about 800 ns after the longest delay.

As a precision pulse generator, the four time intervals define two pulses for applications that require precisely controlled pulse widths. The position and width of each pulse may be programmed from the front panel or via the GPIB. Front panel BNCs provide fast outputs at TTL, NIM, ECL or continuously adjustable levels. These pulses and their complements are available from separate front panel outputs. The outputs may be set to drive either 50-ohm or high-impedance loads.

Fieldtech Heathrow Ltd, Huntavia House, 420 Bath Road, Longford, WEST DRAYTON UB7 0LL.

20 MHZ OSCILLOSCOPE WITH FULL BANDWIDTH AND HIGH SENSITIVITY

The Kenwood CS4025 oscilloscope is a low-cost 20 MHz model with a 80×100 mm (8×10 div.) screen. Its sensitvity is adjustable between 5 mV/div. and 5 V/div. over the full bandwidth; high-sensitive positions of 1 mV/div. and 2 mV/div. are available up to 5 MHz.

The sweep time can be varied from 0.5 µs/div to 0.5 s/div; a maximum sweep speed of 50 ns/div. can be achieved with the use of the ×10 magnifier.

Cross-talk is specified as at least–40 dB for a 1 kHz sine wave.

A useful vertical amplifier signal output (channel 1) is available, which gives an output of 50 mV/div. over a bandwidth of 100 Hz to 10 MHz. This enables, for instance, a frequency counter to be connected for accurate measurement of a waveform at low frequency.

Thurlby-Thandar Ltd, 2 Glebe Road, HUNTINGDON PE18 7DX, Telephone (0480) 412451.

TUNED UHF TV PREAMPLIFIER

based on a design by K. Kraus

For the reception of weak UHF-TV signals a good antenna is, of course, indispensable, but by itself it may not be enough. If that's the case, the preamplifier described in this article may be just what the doctor ordered.

IT is fairly straightforward to couple the antenna to a wideband amplifier, but that could give troubles if the weak signals to be received are close in frequency to strong signals. Even if the amplifier is a good-quality type, the most likely result is a fair dose of cross modulation and all that goes with it. This may be prevented by making the amplifier tunable over a relatively narrow range of frequencies. Even a strong transmitter in a channel adjacent to that of the weak signal can then be suppressed to a fair degree, but the present preamplifier can cope only to some degree if the wanted weak signal is surrounded by a number of strong signals.

The circuit

Basically, the circuit in Fig. 1 consists of two tuned circuits and a dual-gate MOSFET. The signal from the antenna is fed to the MOSFET via a tap on the input circuit, which consists of a 30×3.5 mm stripline. This arrangement ensures that the input impedance is 50–75 Ω. The input circuit is tuned with C1.

MOSFET T1 is arranged as a grounded-source amplifier in which C3 forms the ground connection for HF signals. The drain impedance is formed by the second tuned circuit, which also consists of a stripline, and a small trimming capacitor, C5. The gain of T1 is a maximum at the resonant frequency of the second tuned circuit. Capacitor C4 prevents the DC supply to the MOSFET being short-circuited by the tuned circuit. At the same time, inductor L1 prevents the HF signals being short-circuited by the power supply. The output of the units is taken from a tap on the second tuned circuit to obtain an output impedance of 50–75 Ω.

The DC operating point, determined by the voltage between gate 1 and the source of T1, is set by R2. Owing to the spread of parameters of the MOSFET, the operating point may differ from type to type, and it may therefore be necessary to alter the value of R2 to make certain that the source current is about 10 mA. In the prototype, a value of 150 Ω resulted in a source current of around 12 mA. The value of R2 will be somewhere between 100 Ω and 220 Ω.

Technical data

Frequency range	400–750 MHz*
	400–800 MHz†
Bandwidth at f_c = 500 MHz	492–513 MHz
Gain at 500 MHz	15 dB
Max. attenuation outside pass band	about 40 dB
Output noise (measured with Z_{in} = Z_{out} = 50 Ω)	≤ –80 dBm

* C1, C5 = 1.5–5 pF
† C1, C5 = 0–5 pF

Network R1-C2 ensures that gate 2 is at ground potential for HF signals, so that these signals are processed via gate 1 only.

Power supply

There are various ways of providing power to the amplifier. If it is mounted near the antenna, the supply may be connected via the coaxial antenna feeder cable—see Fig. 2.

The direct voltage is applied to voltage regulator IC1 via inductor L2. The output of the regulator is fed to the cable via the circuit shown in Fig. 2.

If the supply is not via the coaxial cable, L2 may be omitted and an unregulated voltage applied to A.

In case a regulated voltage of 8 V is available, IC1 and C8 may be omitted. The wire bridge between B and C must then be replaced by one between A and B.

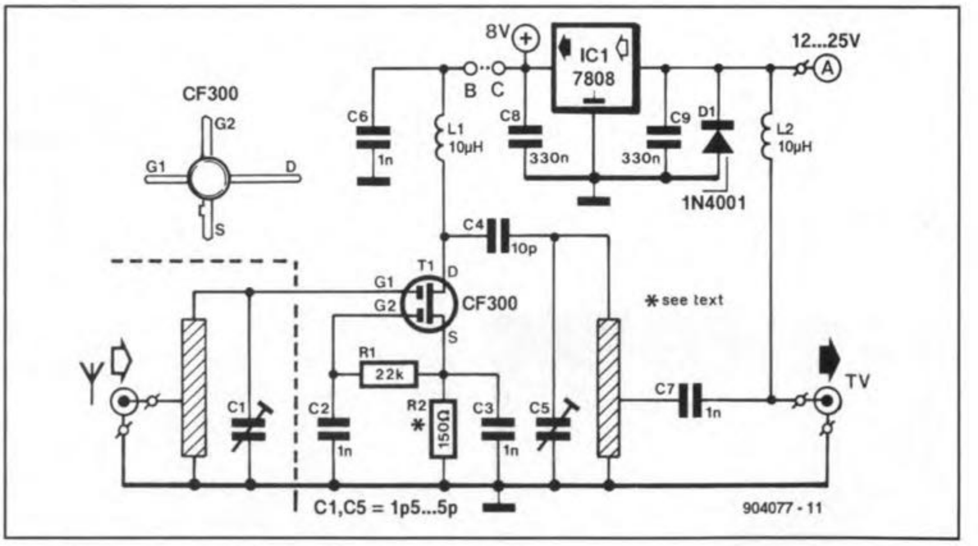


Fig. 1. Circuit diagram of the tuned UHF-TV preamplifier.

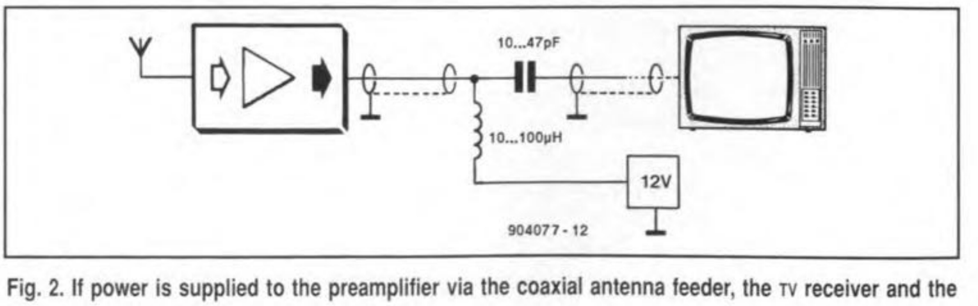
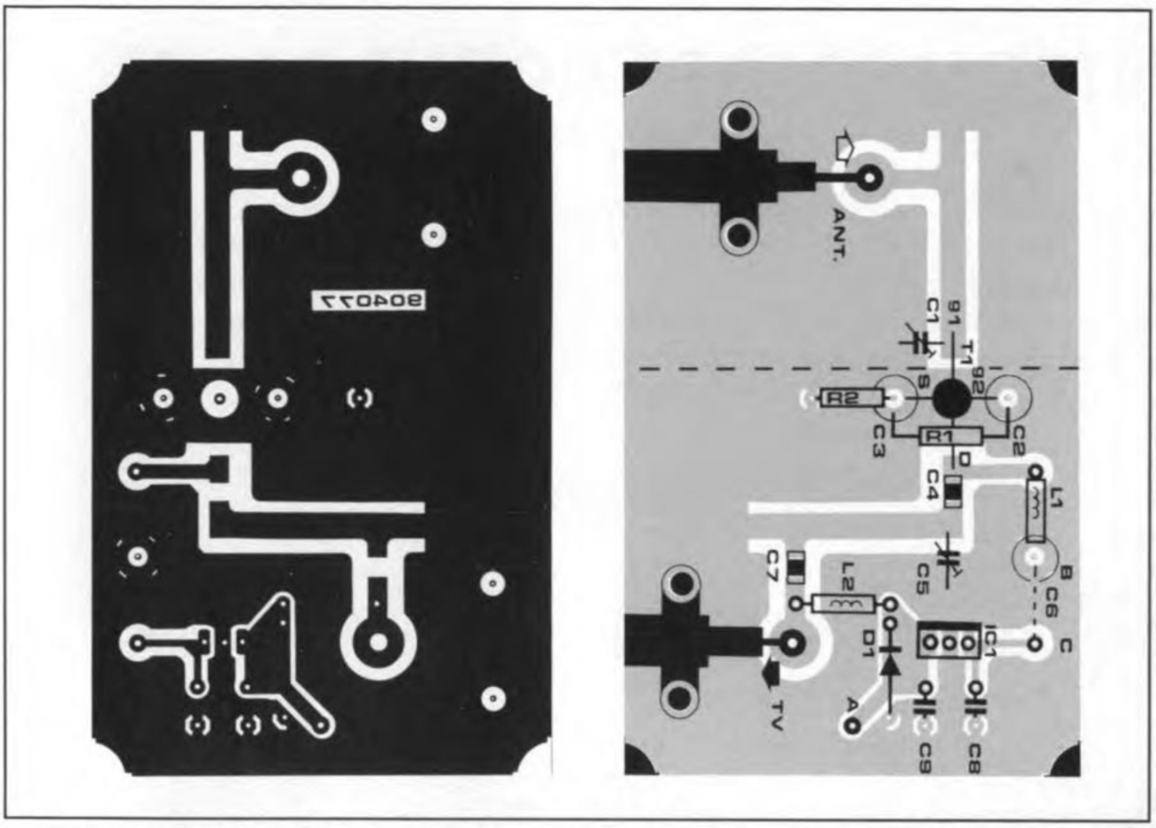


Fig. 2. If power is supplied to the preamplifier via the coaxial antenna feeder, the TV receiver and the power supply must be connected as shown.



PARTS LIST

Resistors:
R1 = 22 k
R2 = 150 Ω (see text)

Capacitors:
C1, C5 = 0–5 pF trimmer (Murata); 1.5–5 pF is permissible (see Technical Data)
C2, C3, C6 = 1 n disk
C4 = 10 p*
C7 = 1 n*
C8, C9 = 330 n

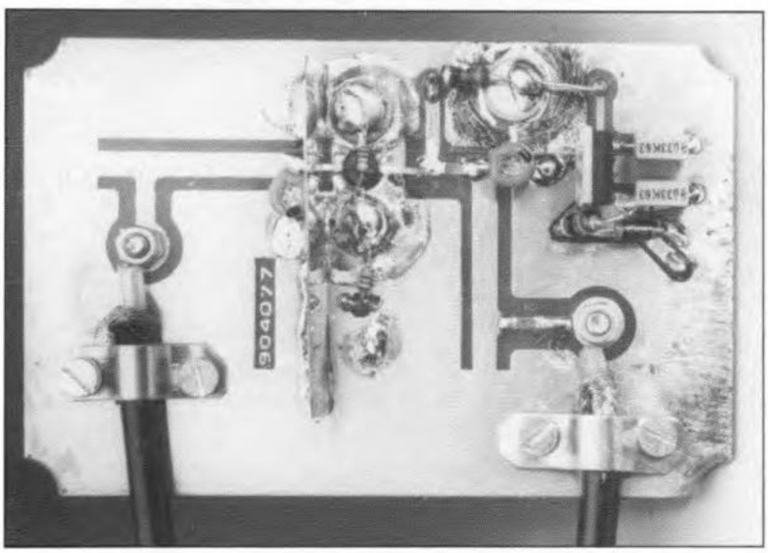
Inductors:
L1, L2 = 10 μH

Semiconductors:
D1 = 1N4001
T1 = CF300 (Telefunken)
IC1 = 7808

Miscellaneous:
PCB Type 904077

* surface mount type

Fig. 3. The printed circuit board for the preamplifier is (or must be) made of epoxy resin. When the board is made, rather than bought ready-made through our Readers' Services, great care should be taken in the dimensioning of the striplines.



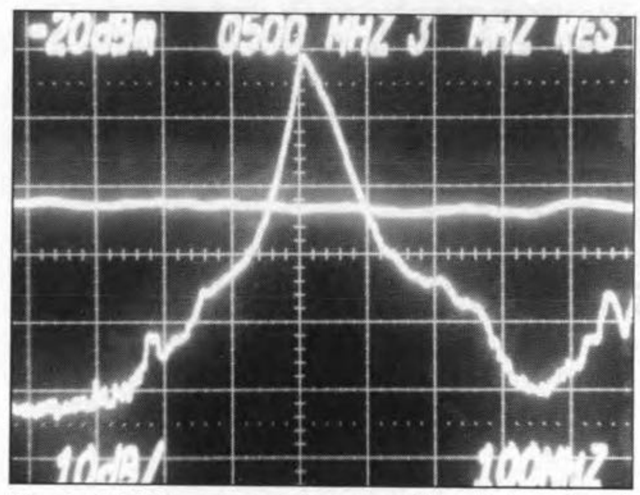


Fig. 4. Selectivity curve of the preamplifier. Horizontal scale 100 MHz/div; vertical scale 10 dB/div.; centre of graph at 500 MHz.

Construction and alignment

The preamplifier must be constructed on the PCB shown in Fig. 3. Note that the component side is also the track side. If you have the board made, make sure that the dimensions of the striplines are exactly right.

Mount disk capacitors C2, C3 and C6, followed by surface-mount capacitors C4 and C7. Next, install the MOSFET between C2 and C3. Solder gate 2 and the source of this transistor to the top terminal of C2 and C3 respectively. Solder gate 1 and the drain to the tracks underneath them.

Solder R1 between C2 and C3 above the MOSFET (although Fig. 3 shows it alongside the transistor).

Next, solder the remaining components in place. Connect R2, L1 and wire bridge B–C (or A–B: see under "Power supply") to the top terminal of the relevant disk capacitor. Mount capacitor C1 as far away as possible from T1, C2 and C3 to ensure that the screen—shown by the dashed line—fits neatly between them.

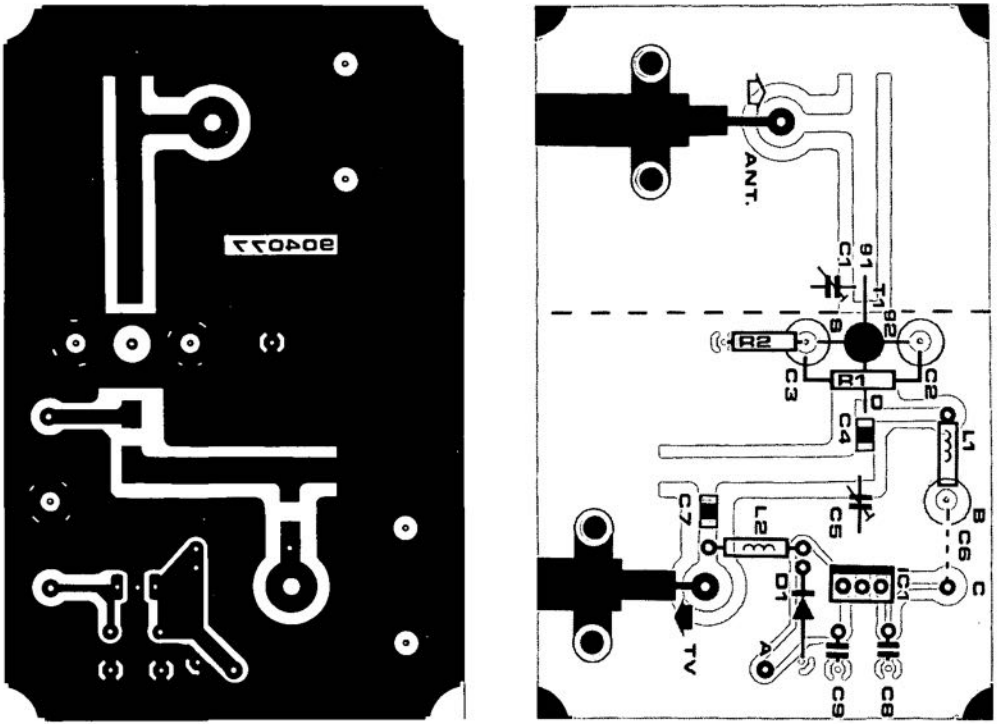
Then, solder a 45×20 mm screen, made of thin tinplate, in the position shown by the dashed line.

Finally, connect the coaxial cables to the relevant terminals and secure them in place with cleats as shown in the photograph of the completed prototype board.

The alignment is pretty straightforward: turn C1 and C5 until the wanted signal is at maximum level and any unwanted ones at minimum level; in other words, until the picture on your TV screen and the associated sound are at an optimum.

If the preamplifier is mounted near the antenna, it may still be aligned as discussed, since the input and output impedance it 'sees' at either end of the feeder cable are equal (at least, if everything is all right).

Finally, a word about the trimming capacitors. The parts list shows two possible types: the 1.5–5 pF is an inexpensive type, but its use makes tuning to the higher channels in the UHF-TV band impossible. If operation in that part of the band is required, there is no choice but to buy the more expensive 0–5 pF Murata type. ■



PARTS LIST

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C7 = 1 n*
C8, C9 = 330 n

Inductors:
L1, L2 = 10 μH

Semiconductors:
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T1 = CF300 (Telefunken)
IC1 = 7808

Miscellaneous:
PCB Type 904077

* surface mount type

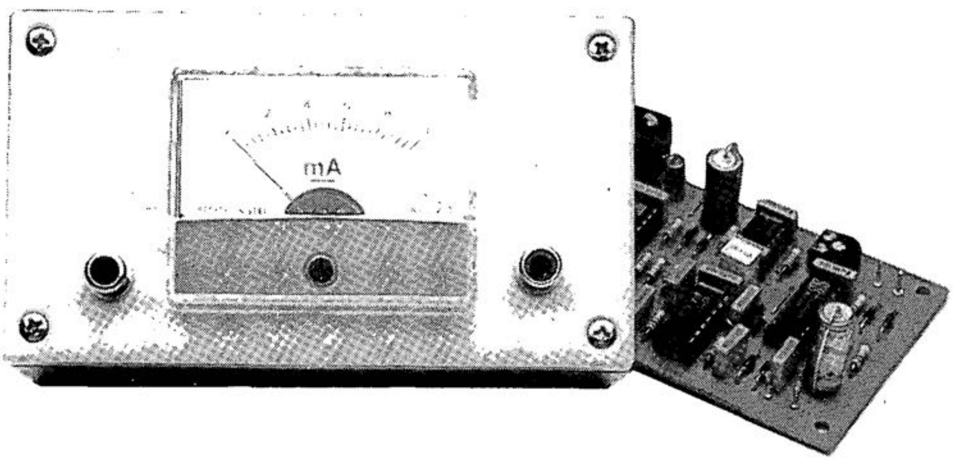
Fig. 3. The printed circuit board for the preamplifier is (or must be) made of epoxy resin. When the board is made, rather than bought ready-made through our Readers' Services, great care should be taken in the dimensioning of the striplines.

INTERMEDIATE PROJECT

A series of projects for the not-so-experienced constructor. Although each article will describe in detail the operation, use, construction and, where relevant, the underlying theory of the project, constructors will, none the less, require an elementary knowledge of electronic engineering. Each project in the series will be based on inexpensive and commonly available parts.

PHASE METER

Phase, phase difference and phase shift are commonly used terms in electronics theory. Where an oscilloscope or a vectorscope are not available, a special test instrument is required to measure phase shift so that it can be checked against the calculated value. Such an instrument is described here.



J. Bareford

THE term phase is used when describing two or more alternating voltages of which the frequency is equal, while their zero-crossings occur at different instants. An alternating voltage crosses the zero line when its instantaneous voltage is nought. An example of two alternating voltages of different phase is shown in the oscilloscope photograph in Fig. 1. The phase difference is about one sixth of the period in this case. The difference is, however, more commonly expressed as an angle because a sine-wave is periodical, so that its period, T, may be described as a circle, or 360°. This means that any part of the period may be expressed in degrees: a period of $0.25T$, for instance, equals 90°, $0.10T$ equals 36°, etc. The phase shift between the two sine-waves in Fig. 1 is, therefore, roughly 60°.

What causes phase shift?

Since the phase difference between two separate voltage sources is hardly ever constant because of small frequency deviations, it makes little sense to design a test instrument for this purpose. The meter described here, however, is intended to locate signal changes that occur in a circuit. Take, for instance, an inverting amplifier. The term 'inverting' implies that the input and output signals are 180° out of phase. In some cases, it is useful to be able to check this.

Transistors and opamps are not the only components that can cause phase shift. Capacitors and inductors introduce a phase shift of 90° between voltage and current. In the case of the capacitor, the voltage lags the current by 90° (see Fig. 2). Inductors have the opposite behaviour: the current lags the voltage by 90°.

Capacitors and inductors are typically used in filters. An example is shown in Fig. 3. In this circuit, one would expect the output voltage to be in phase with the current through the capacitor, i.e., there exists a phase shift of 90° between the input and the output. Unfortunately, this is not so because the phase shift depends on the frequency (the actual relation is an inverse one). The dependency is caused by the ratio of the capacitor's reactance to the resistor value. Since the reactance, x_c, is inversely related to the frequency, it is large with respect to the (fixed) resistor value at low frequencies. Hence, the phase difference is roughly 90° at low frequencies, and 0° at high frequencies.

Phase difference: what it does

In many case, the phase relation between alternating voltages is of little or no importance. There are, however, situations in which the phase relation is all-important. Take the case of two alternating voltages that are to be added. If they are 180° out of phase, the summing operation would result in total cancellation of the two, i.e., an output of 0 V. The equalizing current that flows as a result of this operation is wasted.

In large power plants, phase differences

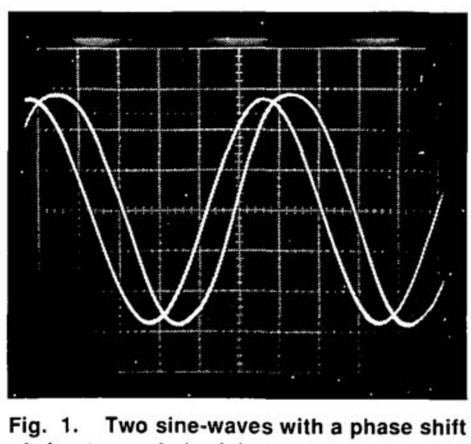


Fig. 1. Two sine-waves with a phase shift of about one sixth of the period, or about 60 degrees.

and the resulting equalizing currents must be kept to a minimum. Generators are, therefore, electrically coupled to maintain synchronicity. Furthermore, many power plants are linked by a special network, which often extends beyond country borders.

Fortunately, problems with phase differences in power plants need not bother us here, and can be safely left to power electricity engineers. The stereo equipment in your living room may, therefore, be a better example to illustrate the importance of steady phase relations. Since any amplifier in the system is bound to contain a filter of some kind, it may be expected that there is some phase shift between the input source, e.g., a CD player, and the output, the loudspeaker. Evidently, it is desirable for this shift to be roughly the same over the full frequency range of the signal. If this is not so, the sound is not perfect, particularly when the phase differences between the left and the right channel is large.

A final example of a circuit in which phase shift is all-important is the oscillator, which functions only when the amplified output signal is fed back to the input at a phase shift of 0°.

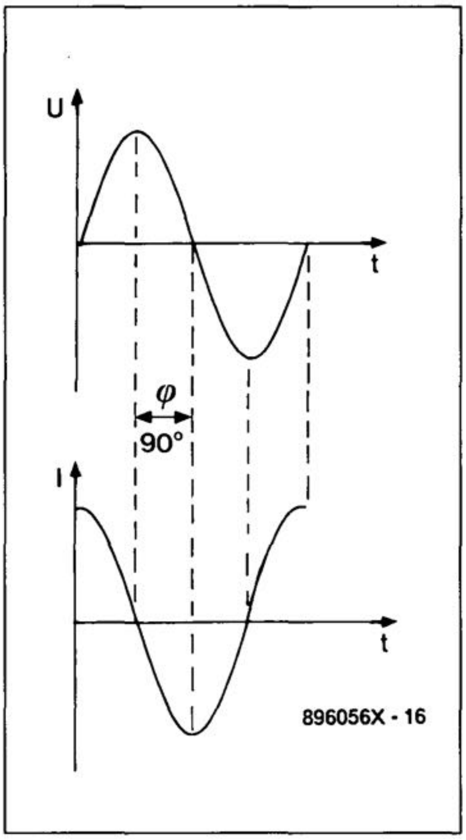


Fig. 2. The voltage on a capacitor lags the current by 90 degrees.

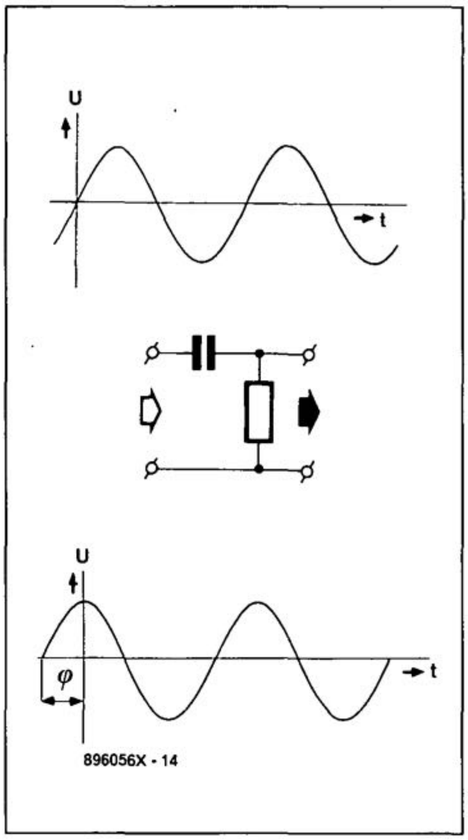


Fig. 3. Phase shift introduced by an *R-C* network

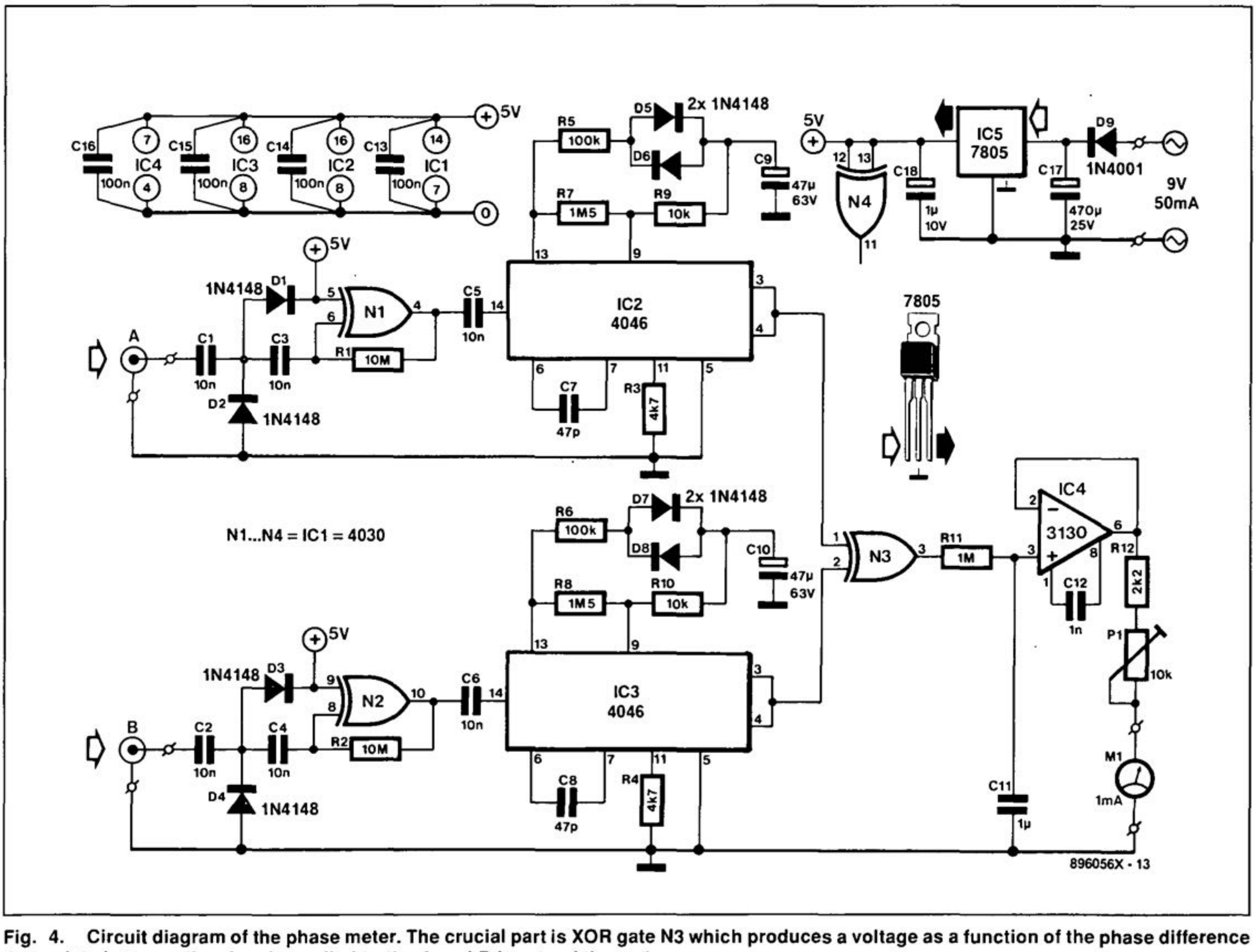


Fig. 4. Circuit diagram of the phase meter. The crucial part is XOR gate N3 which produces a voltage as a function of the phase difference that exists between the signals applied to the A and B inputs of the meter.

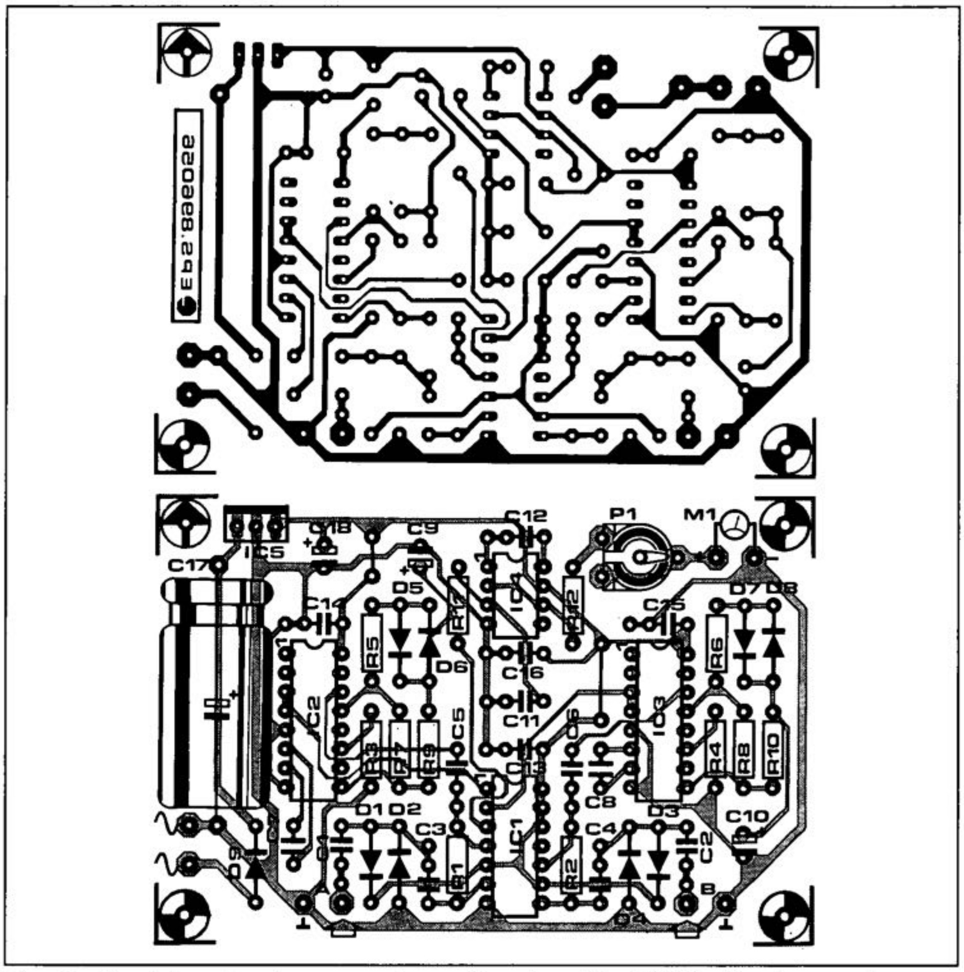


Fig. 5. Track layout and component mounting plan of the PCB for the phase meter.

COMPONENTS LIST

Resistors:

2	10MΩ	R1;R2
2	4kΩ7	R3;R4
2	100kΩ	R5;R6
2	1MΩ5	R7;R8
2	10kΩ	R9;R10
1	1MΩ	R11
1	2kΩ2	R12
1	10kΩ preset H	P1

Capacitors:

6	10nF	C1 – C6
2	47pF	C7;C8
2	47μF 63V radial	C9;C10
1	1μF	C11
1	1nF	C12
4	100nF	C13 – C16
1	470μF 25V	C17
1	1μF 10V radial	C18

Semiconductors:

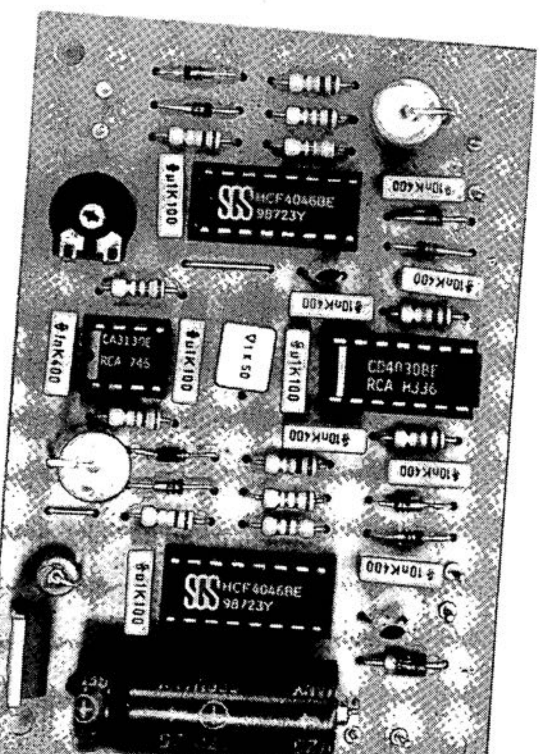
8	1N4148	D1 – D8
1	1N4001	D9
1	4030B	IC1
2	4046B	IC2;IC3
1	CA3130	IC4
1	7805	IC5

Miscellaneous:

1	1 mA moving-coil meter	M1

The phase meter

As illustrated in Fig. 1, a two-channel oscilloscope is perfect for measuring phase differences. Unfortunately, such an instrument is not available to every one, and this is where a special phase meter comes in.



The circuit diagram of the two-channel phase meter is shown in Fig. 4. The capacitors at inputs A and B block any d.c. components in the signals applied to the circuit. Diodes D_1 – D_4 protect amplifiers N_1 and N_2 against negative or too high voltages. The amplified signals are subsequently fed to IC_2 and IC_3, two phase-locked loops (PLLs) that track the input frequencies and convert them to digital signals with a fixed duty factor of 0.5 and a swing of 5 Vpp. The low-pass filters of these PLLs are special types to enable the 4046s to keep track of fast frequency changes in the input signals. This important characteristic is provided primarilyby the diodes in in the network connectd to pins 13 and 9.

The actual phase comparator is formed by a single XOR (exclusive-OR) gate, N_3. The output signal of the gate is averaged by integrator network R_{11}-C_{11}. The larger the phase difference between the two input signals, the larger the voltage supplied by buffer IC_4. Moving-coil meter M_1, resistor R_{12} and preset P_1 together form a meter that indicates the phase shift. Since the voltage indicated by the meter lies between 0 V and the positive supply level, the circuit must be powered from a regulated supply. A three-terminal fixed voltage regulator Type 7805 is used here for this purpose. The minimum unregulated or alternating input voltage is 9 V.

Construction and setting up

The circuit is best constructed on the printed-circuit board shown in Fig. 5. First, fit the two wire links so that they are not forgotten later. Pay attention to the orientation of the polarized components (diodes, transistors, ICs and electrolytic capacitors). Fit IC sockets for all ICs.

To align the meter, first remove IC_2 and IC_3 from their sockets. Connect pin 3 of the socket for IC_2 to pin 5 (ground) via a 10 kΩ resistor. Next, connect pin 3 of the socket for IC_3 to pin 16 (+5 V) via a 10 kΩ resistor. Switch on and check that pin 1 of N_3 is at about 0 V, and pin 2 at about +5 V. The output of the XOR gate, pin 3, should be at +5 V. Since this represents the maximum phase shift that can be measured between the two channels, adjust preset P_1 until a meter indication of 180° is obtained.

Remove the two 10-kΩ resistors and fit the two 4046s. At this stage, the circuit is ready for fitting into an enclosure.

Although the indication range of the meter is, in principle, limited to 180°, it is possible to measure greater phase shifts. A shift of 270°, for instance, is indicated as 90° because the meter can not detect which channel, A or B, was high first. ■

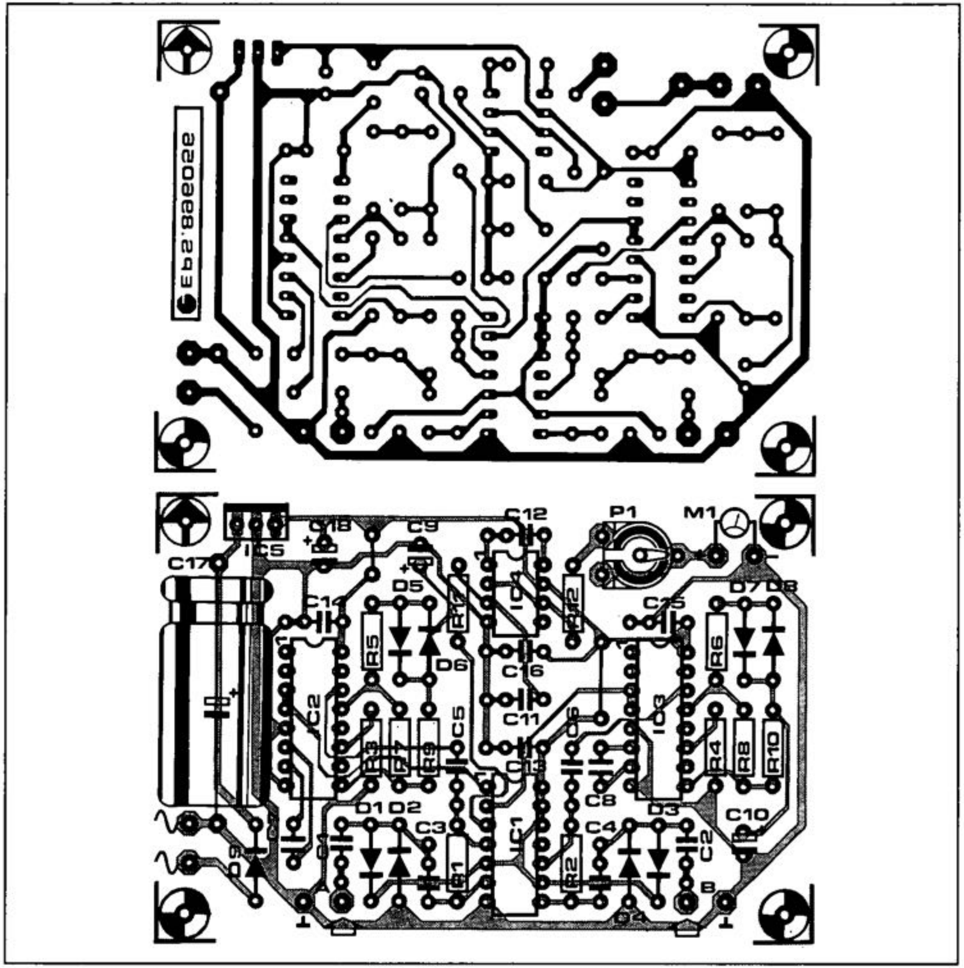


Fig. 5. Track layout and component mounting plan of the PCB for the phase meter.

COMPONENTS LIST

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2	10kΩ	R9;R10
1	1MΩ	R11
1	2kΩ2	R12
1	10kΩ preset H	P1

Capacitors:

6	10nF	C1 – C6
2	47pF	C7;C8
2	47µF 63V radial	C9;C10
1	1µF	C11
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4	100nF	C13 – C16
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1	1µF 10V radial	C18

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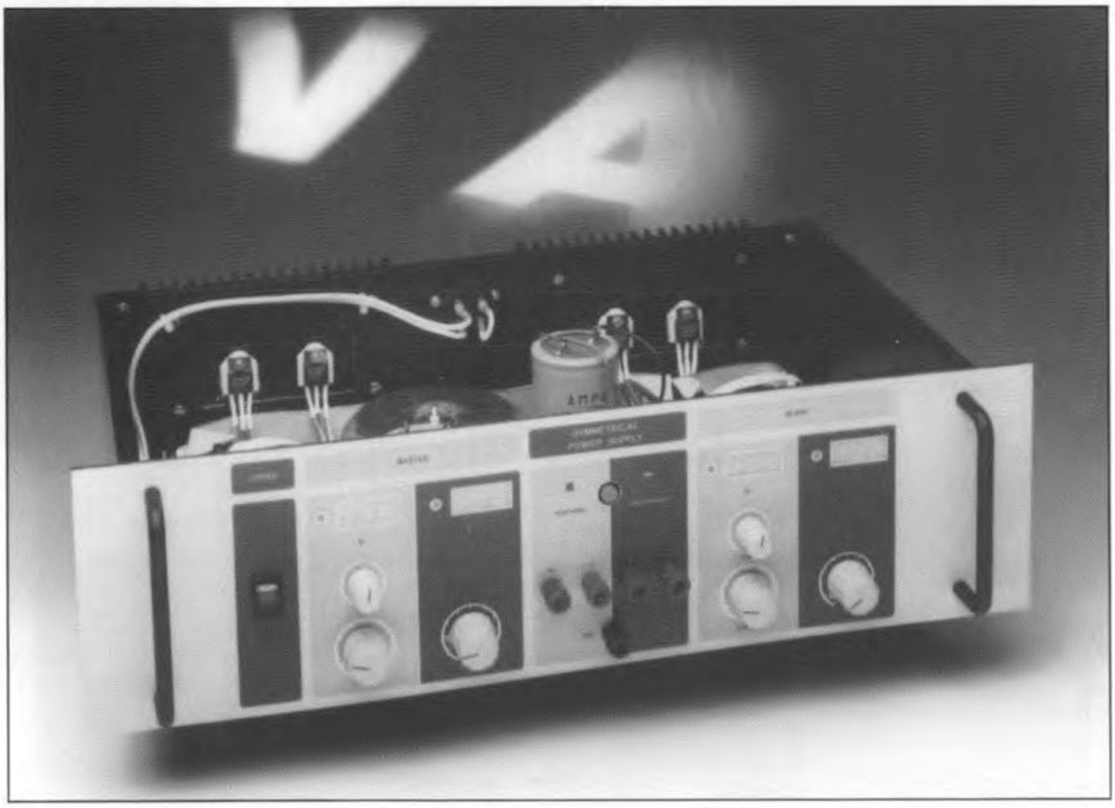
8	1N4148	D1 – D8
1	1N4001	D9
1	4030B	IC1
2	4046B	IC2;IC3
1	CA3130	IC4
1	7805	IC5

Miscellaneous:

1	1 mA moving-coil meter	M1

400-WATT LABORATORY POWER SUPPLY



PART 1: CIRCUIT DESCRIPTION

Here is an all-purpose d.c. power supply for symmetrical as well as asymmetrical use, and capable of supplying high output currents and voltages. An all-analogue design based on discrete parts only, this 400-watt PSU deserves a prominent place on your work bench.

G. Boddington

The problem with power supplies in an electronics laboratory or workshop is that their application range is often limited because of their specifications. Any one who has been involved in practical electronics will admit that finding a suitable power supply for a particular test is not at all easy, when none of the available ones (say ±15 V/2 A, 0–60 V/100 mA and 5 V/10 A types) seem to be fully geared to the job. Obviously, what is needed is a supply that combines the most frequently used voltage and current ratings, both symmetrical and asymmetrical, while offering a properly operating overload protection.

Although the user manual with many an inexpensive, ready-made power supply will confidently inform you that the power transistors are protected against overloads, this type of protection has an inherent disadvantage. True, the supply will happily supply the maximum output current at the maximum output voltage, but it will typically shut down the moment the voltage is reduced just one volt or so. The reason is clear: the overload protection is actuated because the extra dissipation, which is the product of the output current and the voltage difference across the series transistors, exceeds the cooling capacity of the heat-sink, or the power rating of the (expensive) series transistors.

The present power supply puts an end to

MAIN SPECIFICATIONS

- **Mode: Single**
  - one adjustable power supply with current and voltage controls.
  - Output: 0 – 40 V at 0 – 5 A
- **Mode: Independent**
  - two identical, electrically separated, power supplies.
  - Outputs: 2 × 0 –4 0 V at 2 × 0 – 5 A
- **Mode: Tracking**
  - two identical, series connected, power supplies.
  - Outputs: ±0 –± 40 V at 0 –5 A; 0 – 80 V at 0 – 5 A
  - Voltage and current of *slave* follow *master*.
- **Mode: Parallel**
  - two identical, parallel connected, power supplies.
  - Outputs: 0.6 – 39.4 V at 0 – 10 A
- **Maximum output voltage:** **40 V (at full load)**; **48 V (no load)**
- **Maximum output current:** **5 A**
- **Ripple:** **10 mV (no load)**; **50 mV (at full load)**
- **Voltage difference in tracking mode:** **50 mV**

these problems. It can be set up to supply either 2 × 40 V/2 × 0 – 5 A, ±0 – 40 V/0 – 5 A, or 0 – 80 V/0 – 5 A, and is capable of supplying the maximum output current at low voltage settings. Special ICs or microprocessors are not used: just straightforward analogue electronics based on readily available components. The result is a power supply with an excellent price/performance ratio.

Block diagram

The instrument consists of two identical, electrically isolated, power supplies, which may be connected in a number of ways to give different operating modes. The block diagram in Fig. 1 shows relatively many functional blocks, which together form three partly 'interwoven' regulating circuits. The first of these, the outer circuit, is a transformer preregulator that serves to keep the voltage drop across the series transistors (T4-T5) constant at about 10 V, so that the maximum dissipation remains smaller than 50 W (or 25 W per transistor). The other two regulation circuits are for the output voltage (*U*) and current (*I*). These circuits are almost identical, the only difference being that the current control obtains its control information from a series resistor, and the voltage control from a potential divider fitted across the output terminals. In contrast to the transformer preregulation, the *U* and *I* control circuits allow the range of the regulating action to be adjusted manually. Interestingly, the series transistors, T4 and T5, function in all three regulation circuits.

The block diagram shows a second power supply, which provides auxiliary ±12 V rails for use in the main circuit. The ground line of this symmetrical supply is connected to the positive output terminal of the main supply. This means that all references to '+12 V' and '–12 V' in the following text, and in the circuit diagram, are actually '+12 V and –12 V with reference to the positive output terminal'. The auxiliary power supply also functions as a voltage reference.

Finally, the block marked 'current limit' stands for a circuit that keeps the output current of each supply below 5 A. This circuit may be fitted with an optional temperature monitor to prevent overheating.

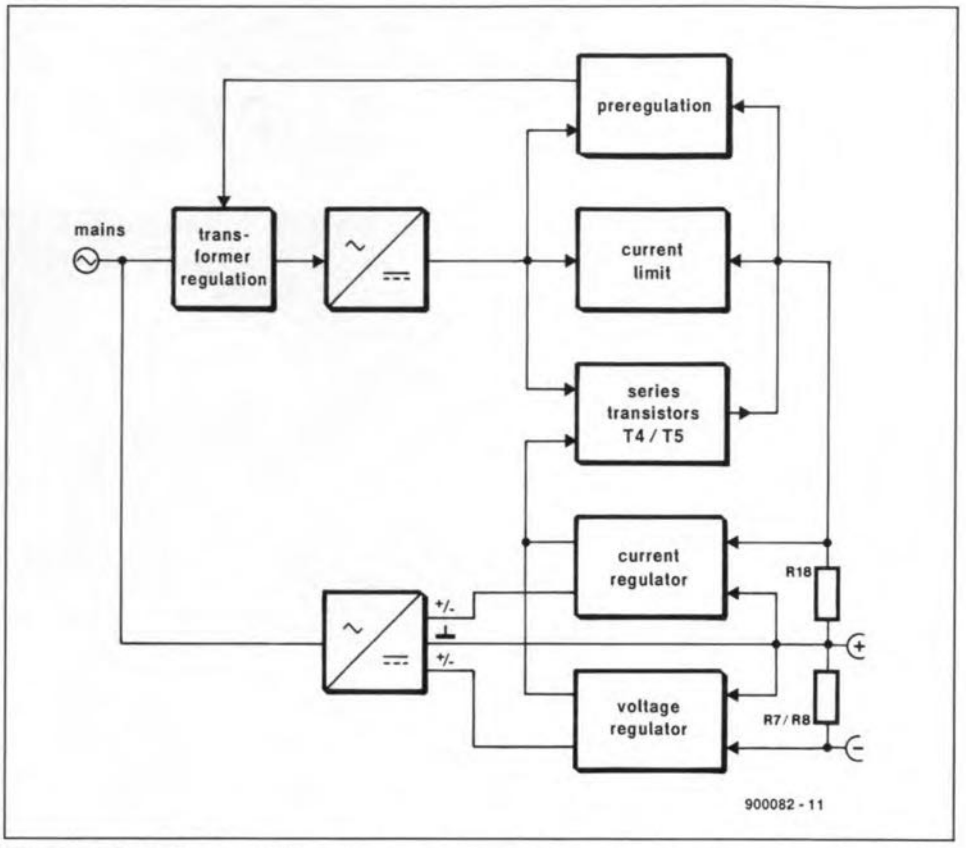


Fig. 1. Block diagram of the power supply. The design is based on three interactive control circuits: (1) transformer preregulation, (2) current control and (3) voltage control.

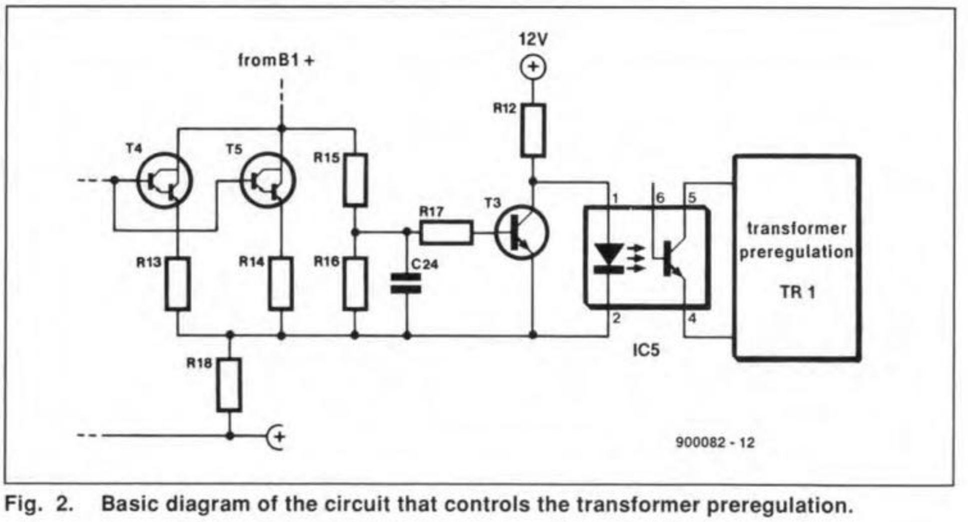


Fig. 2. Basic diagram of the circuit that controls the transformer preregulation.

The preregulation circuit

The basic operation of the preregulation circuit is best explained with reference to Fig. 2. The current flows from the positive connection of the bridge rectifier to the positive output terminal via two parallel-connected darlington transistors, T4 and T5, and resistors R13, R14 and R18. The regulation circuit tries to maintain a constant drop of 10 V across the series transistors and their emitter resistors. Transistor T3 is driven via potential divider R15-R16 and network C24-R17. The network introduces a small delay to eliminate the effect of noise spikes in the preregulation. The current through the LED in optocoupler IC5 is inversely proportional to the voltage across R15-R16.

The power fed to mains-connected ohmic loads is relatively simple to control. Usually, an adjustable *R-C* network connected across the mains terminals supplies the trigger voltage for a triac. The timing of the trigger (or firing-) pulse with respect to the start of the half-cycle is determined by the *R-C* delay. After being fired, the triac conducts until the mains voltage drops to a level below the minimum hold current. This happens close to the zero-crossing. The triac remains blocked until it receives another trigger pulse at a particular phase angle during the next half-cycle of the mains voltage. The current supplied to the load is inversely related to the phase angle, i.e., to the delay of the trigger pulse following the zero-crossing. This principle of phase-angle control works as long as voltage and current are in phase, i.e., as long as the load is a pure resistance.

Unfortunately, the mains transformer in the power supply forms an inductive rather than an ohmic load, so that the mains voltage and the load current are out of phase. Hence, a simple 'dimmer' with conventional triac control as described above will not do as a preregulation circuit. With an inductive load, it may happen that although the instantaneous voltage is high enough to fire the triac, there is no current to 'hold' the device. Therefore, the firing may take place only when the load current is sufficiently high to keep the triac conductive. However, since the load current in a power supply is variable, the phase shift between voltage and current is also variable. This means that the width of the trigger pulse rather than the position must be controlled. If the pulse were simply shifted, the result would be an asymmetrical output current with a high d.c. component, causing rapid saturation of the transformer winding. When the pulse is stretched, however, due care must be taken to prevent it from extending over the zero crossing of the mains voltage.

The circuit section in Fig. 3 stretches the first pulse by means of pulse sequence triggering, an approach which is particularly suited to applications with load currents that are prone to variation. The *R-C* network connected between the live and neutral lines of the mains serves to delay the trigger instant. The network consists of C1, potential divider R29-P1-R30 (branch 1), series resistor R31 and bridge rectifier D20–D23 (branch 2). The combination of the bridge rectifier and the optocoupler it powers simply forms an adjustable resistor for alternating voltages, so that both branches have the same function: making the trigger delay, ϕ, variable (see Fig. 4a). The basic delay is determined by P1.

When the power supply is switched on, C1 is charged. When the trip voltage of the diac is reached, both Di1 and Tri1 are fired. When a trigger current flows from C1 to Tri1, resistor R32 drops a voltage which is high enough to trigger a smaller triac, Tri2. The result is that the discharge time is no longer determined by the two branches, but by (R33+R29)C1. When C1 can no longer supply the hold current for Tri2—which happens fairly quickly because of the small resistors R33 and R29—the triac blocks and C1 charges again. This sequence is repeated until just before the zero-crossing, when the mains voltage can no longer charge C1 (see Fig. 4a). The waveform across the thyristor is shown as a dashed line in Fig. 4b. Figure 4c, finally, shows the waveform of the current shifted by an angle φ as a dashed line, and the waveform produced by the dimmer as a solid line. Asymmetry of the waveform occurs during the first half-cycle only. The triac conducts up to instant 'B', when the load current falls to zero.

The function of the remaining parts in this section of the circuit is quickly explained: the zener diodes limit the voltage across Tri2 to about 66 V whilst providing a stable reference voltage for the trigger circuit. Components D12, D13, D14, R27 and R28 ensure that C1 discharges during the zero-crossing. Inductor L1 serves to eliminate current surges and thus prevent HF interference. Network C2-R34 short-circuits spikes generated by the switching sequences, and so prevents erroneous triggering.

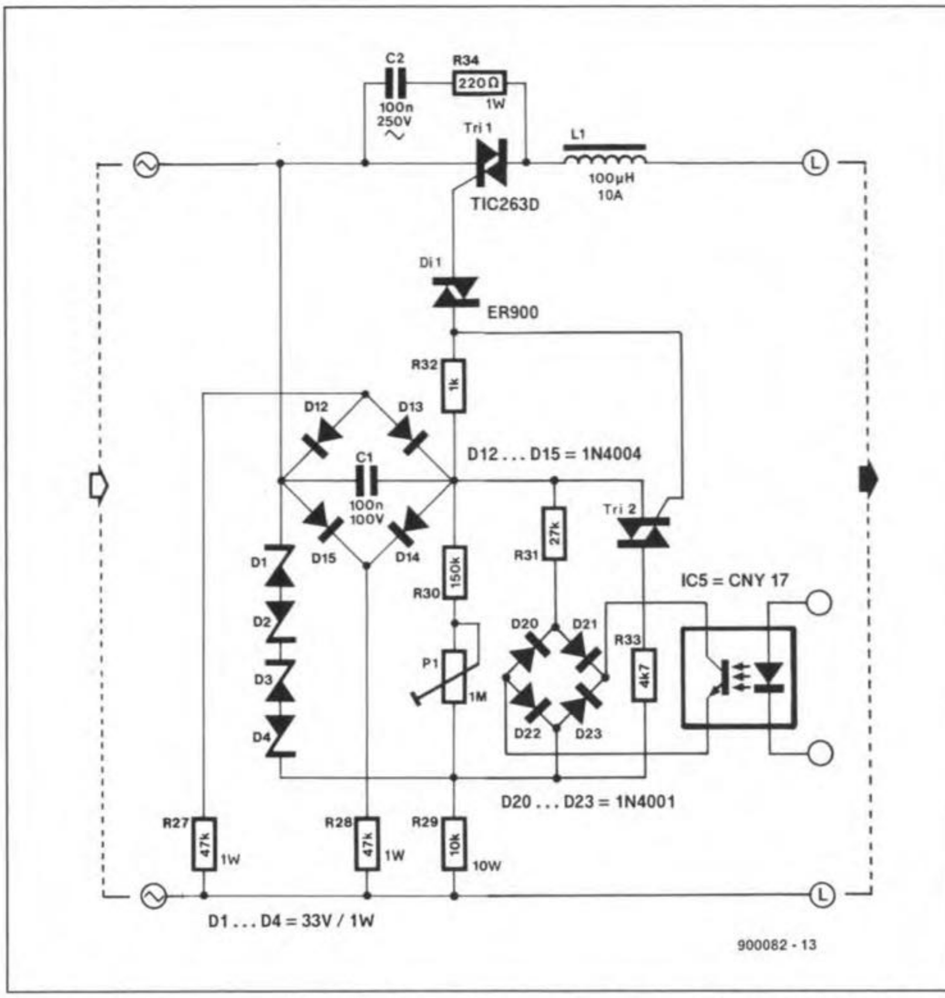
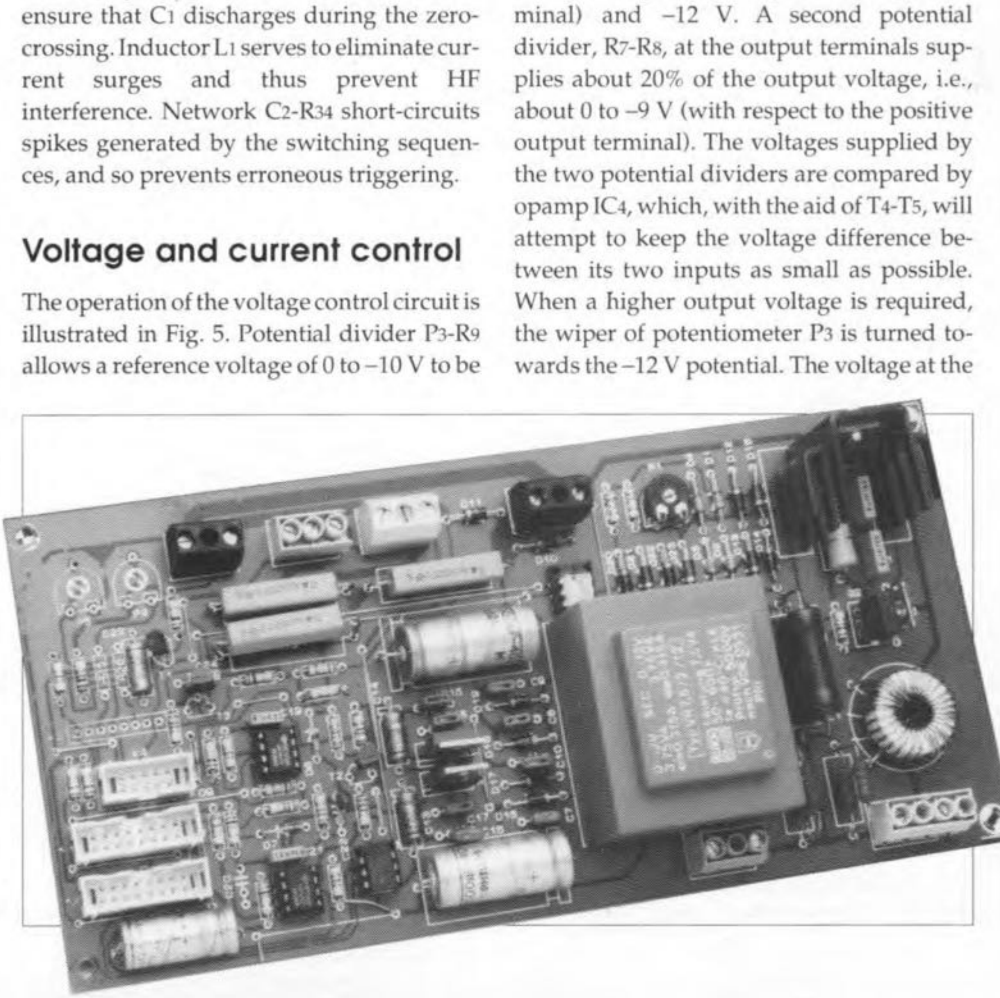


Fig. 3. Circuit diagram of the transformer dimmer. The trigger delay is controlled by the circuit in Fig. 2 via an optocoupler, IC5, and an adjustable bridge rectifier, D20–D23.

Voltage and current control

The operation of the voltage control circuit is illustrated in Fig. 5. Potential divider P3-R9 allows a reference voltage of 0 to –10 V to be set between ground (the positive output terminal) and –12 V. A second potential divider, R7-R8, at the output terminals supplies about 20% of the output voltage, i.e., about 0 to –9 V (with respect to the positive output terminal). The voltages supplied by the two potential dividers are compared by opamp IC4, which, with the aid of T4-T5, will attempt to keep the voltage difference between its two inputs as small as possible. When a higher output voltage is required, the wiper of potentiometer P3 is turned towards the –12 V potential. The voltage at the



non-inverting input of IC4 drops, so that the output voltage of the opamp rises. Conversely, when a lower output voltage is set either by the user turning P3, or by the actuation of the voltage limiting circuit, the inverting input is at a higher potential than the non-inverting input, so that the opamp output voltage drops.

The current control circuit (Fig. 6) operates in a similar manner. Like IC4, opamp IC3 will attempt to keep its output voltage at 0 V. The main difference with the voltage control circuit, however, is that the reference voltage for the opamp (applied to the non-inverting input) is permanently grounded via R1, while the current is measured as the drop (max. 1.1 V) across series resistor R18. Potential divider P2-R3 is arranged so that its junction carries a voltage between –1.1 V and +1.1 V with respect to the positive output terminal. When no current flows through R18, the positive side of P2 is at ground potential. When P2 is advanced to the 5-A position, i.e., to its full resistance of 2.2 kΩ, the inverting input of IC3 is at a voltage of –1.1 V. Consequently, the voltage at the opamp output rises.

When a current of 5 A flows, R18 supplies 1.1 V. When P2 is turned to the other extreme position (i.e., a resistance of 0 Ω), the voltage at the inverting input is higher than that at the non-inverting input, so that the opamp output voltage drops.

As shown by Figs. 5 and 6, and also by the complete circuit diagram in Fig. 7, the anodes of D8 and D24 share a common connection, R23, where the opamp outputs of the current and voltage control circuits are joined. This means that the opamp that supplies the lower output voltage determines the base voltage of the current booster, T4-T5. Resistor R23 serves to hold the bases of T4-T5 at about +11.5 V. Diodes D7 and D9 decouple the opamp outputs, preventing current from flowing between them. One of the series-connected LEDs lights when the voltage at

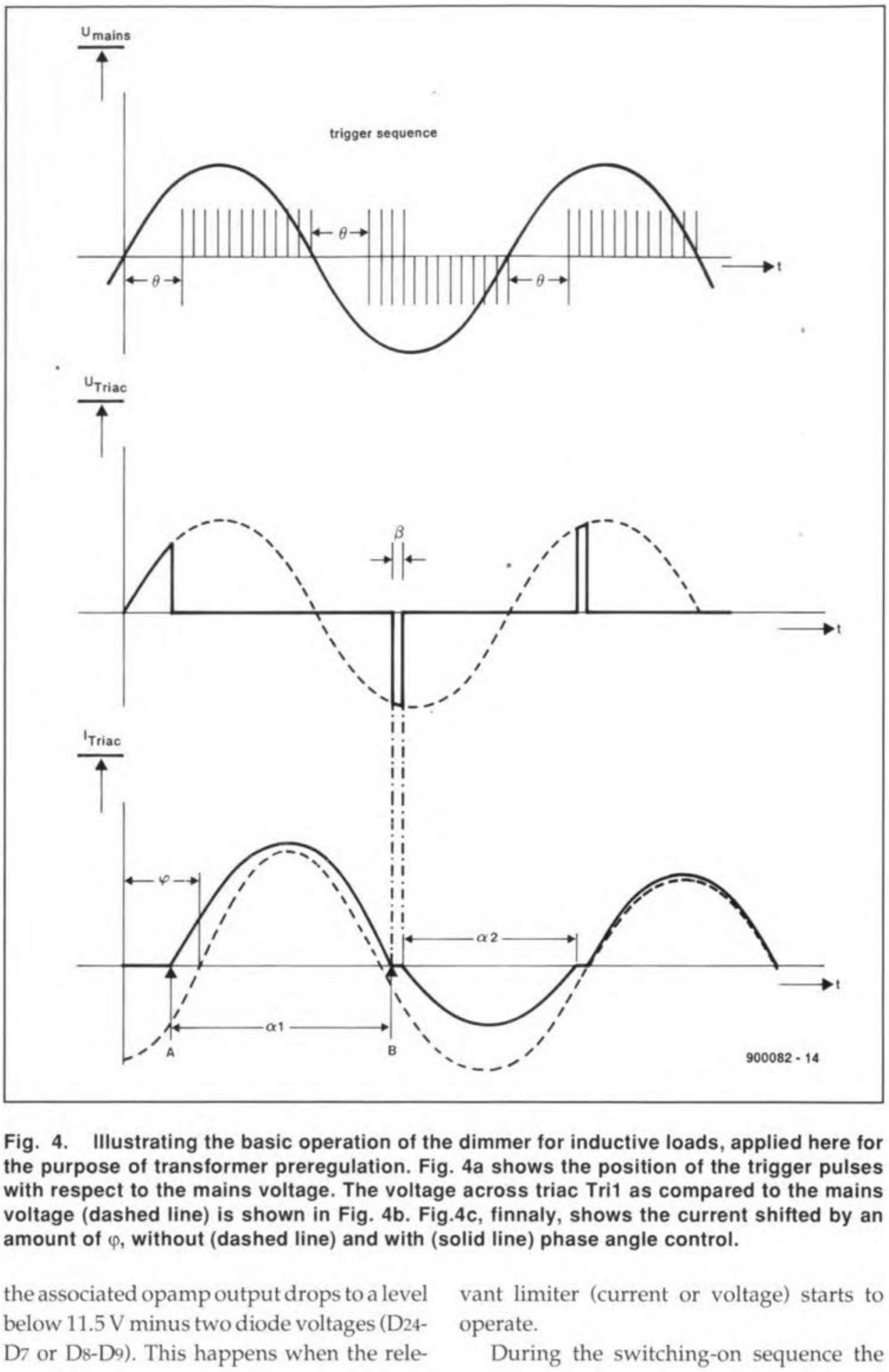


Fig. 4. Illustrating the basic operation of the dimmer for inductive loads, applied here for the purpose of transformer preregulation. Fig. 4a shows the position of the trigger pulses with respect to the mains voltage. The voltage across triac Tri1 as compared to the mains voltage (dashed line) is shown in Fig. 4b. Fig.4c, finnaly, shows the current shifted by an amount of φ, without (dashed line) and with (solid line) phase angle control.

the associated opamp output drops to a level below 11.5 V minus two diode voltages (D24-D7 or D8-D9). This happens when the relevant limiter (current or voltage) starts to operate.

During the switching-on sequence the

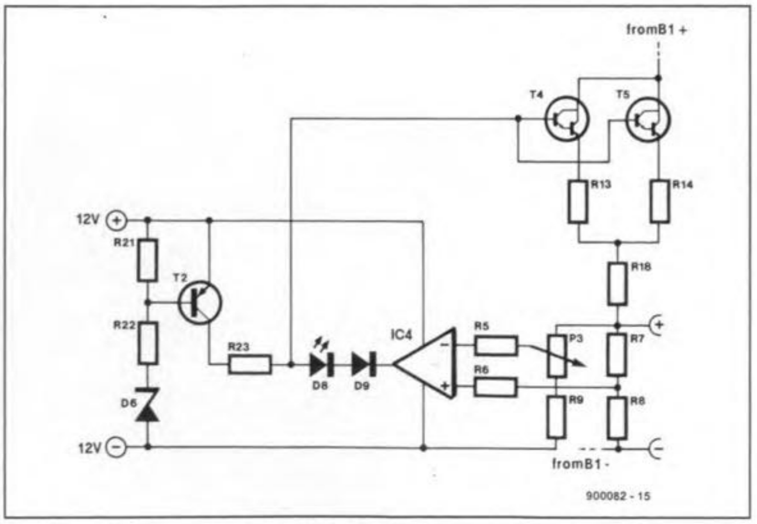


Fig. 5. Basic voltage control circuit.

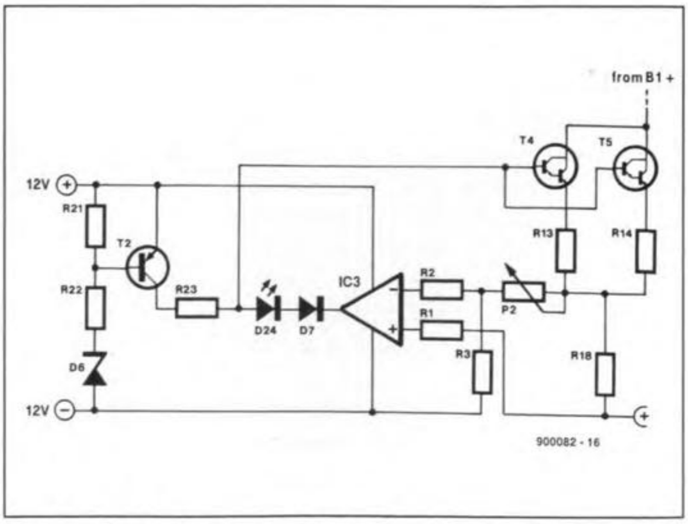


Fig. 6. Basic current control circuit.

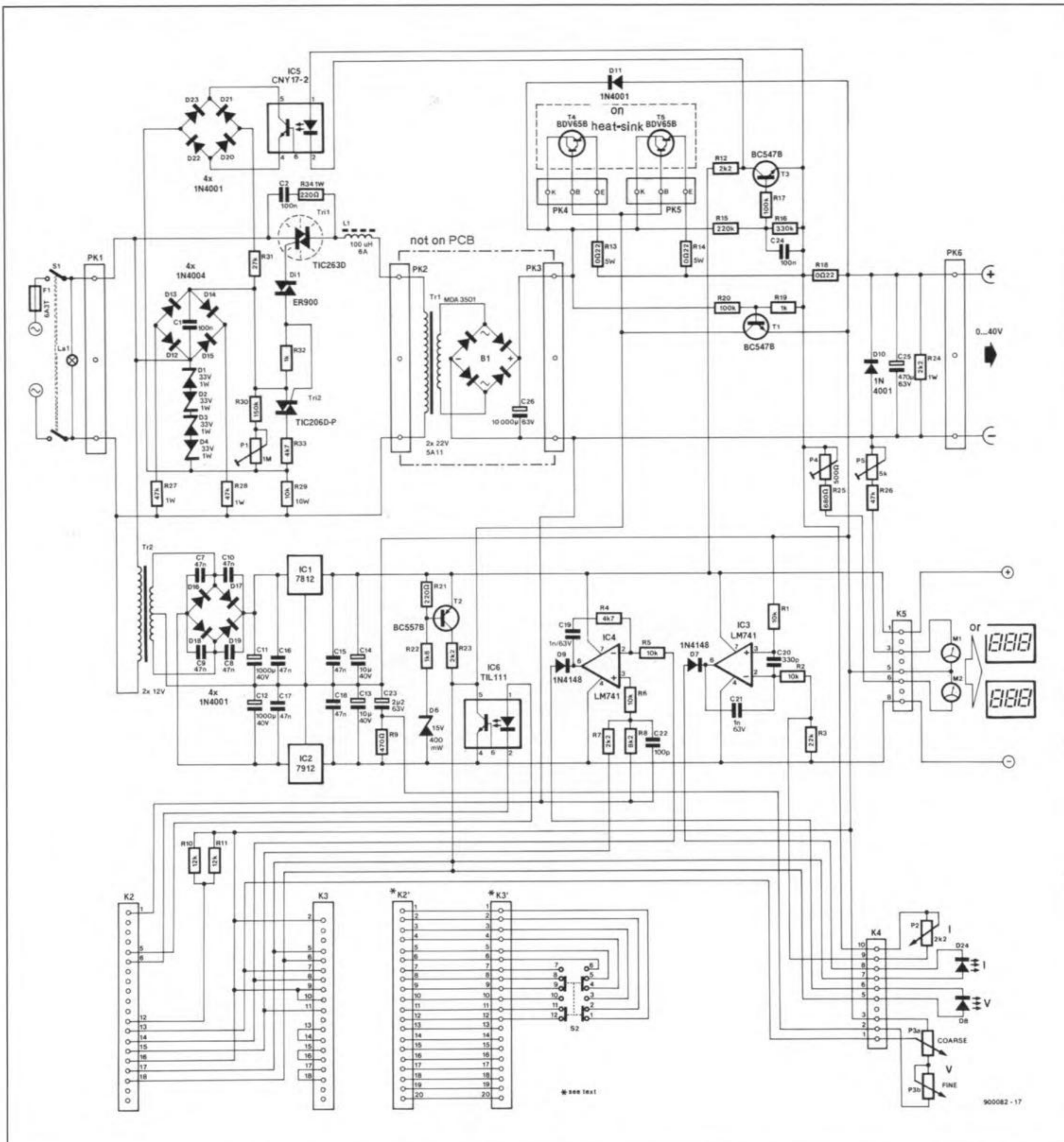


Fig. 7. Circuit diagram ofthe laboratory power supply. Two of these circuits are required for the parallel, series and symmetrical modes.

circuit around T2 keeps the series transistors off until the zener voltage of D6 is reached. This happens when the negative supply voltage of the opamp is sufficiently high. In this way, the voltage peak at switch-on is limited to about 2.5 V above the set output voltage, which is available after a few milliseconds. Although the switch-on peak is not likely to cause damage to most equipment powered by the supply, it is recommended to first switch on the PSU and then connect the load.

The current limiting function of the PSU is provided by the circuit in Fig. 8. As long as the pre-regulation circuit operates correctly, there exists a constant voltage difference across T4-R13 and T5-R14. In a fault condition of any kind (overcurrent, overvoltage), T1 is switched on via potential divider R19-R20. This reduces the base voltage of the darlington transistors, so that the output current is limited. To implement a combined current/temperature overload function, replace resistor R20 by a 100-kΩ NTC (negative temperature coefficient) resistor which is bolted on to the heat-sink, close to T4-T5. ❑

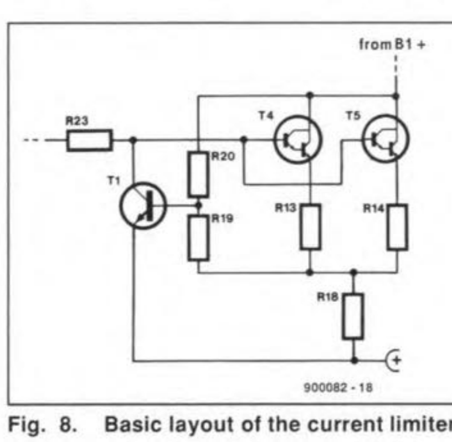


Fig. 8. Basic layout of the current limiter.

CTIONS CORRECTIONS CORRE

400-watt laboratory power supply

October 1989 and November 1990

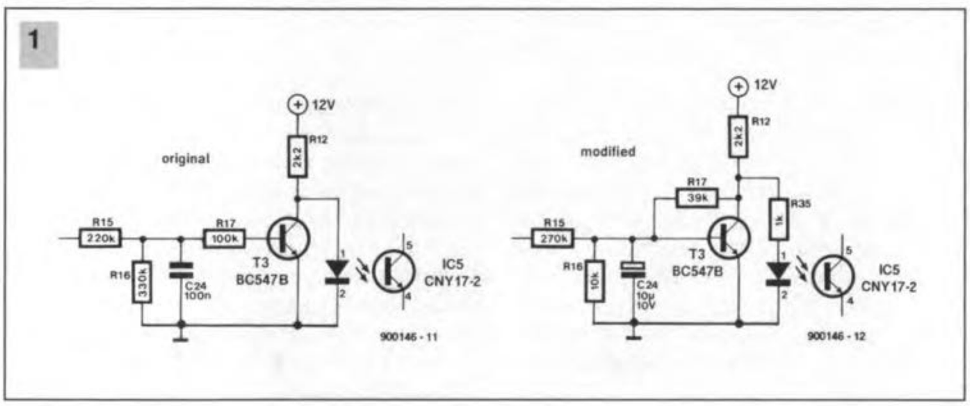
A number of constructors of this popular project have brought the following problems to our attention.

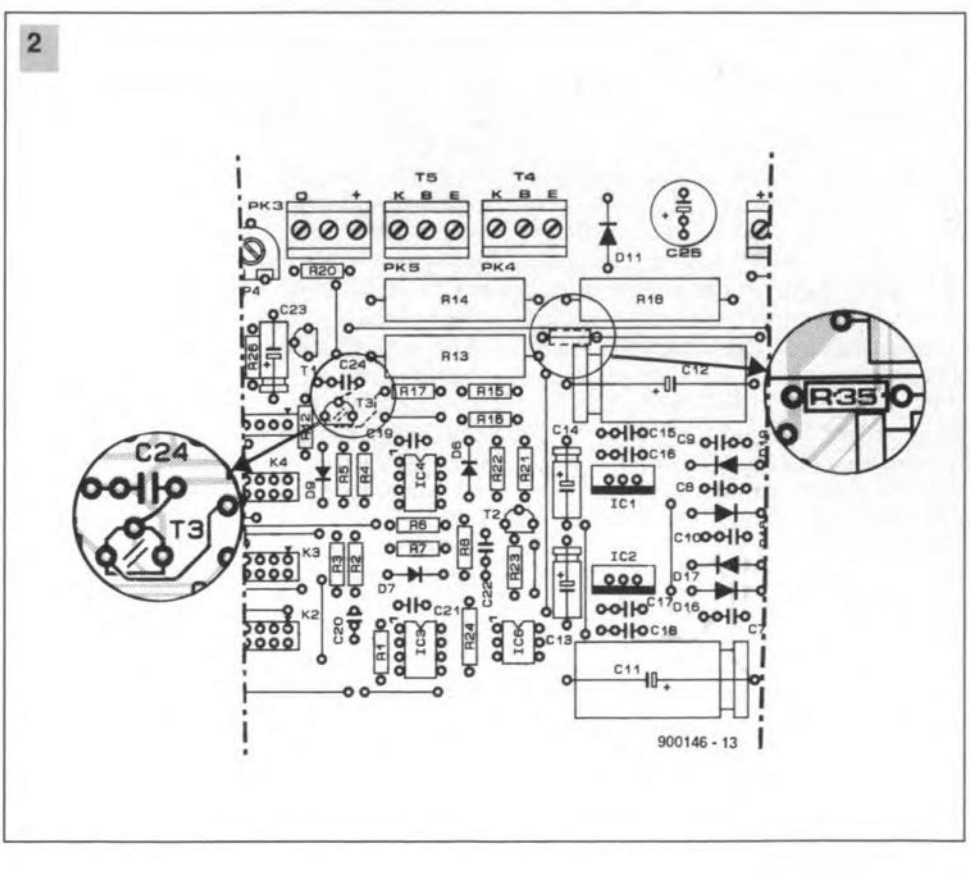
1. The onset point of the current limit circuit lies at about 3 A, which is too low. Solve this problem by replacing T_1 with a Type BC517 darlington transistor, and R_{20} with a 82kΩ resistor.

2. Depending on the current transfer ratio of the optocoupler used, the transformer produces ticking noises. This effect, which is caused by overshoot in the pre-regulation circuit, may be traced with the aid of an oscilloscope monitoring the voltage across C_{26} at a moderate load current. The capacitor must be charged at each cycle of the mains frequency, and not once every five cycles. The problem is best solved by reducing the amplification of the regulation circuit. Replace R_{17} with a 39 kΩ resistor, and create feedback by fitting it between the base and the collector of T_3. Also add a resistor in series with the optocoupler. These two changes are illustrated in Figs. 1 and 2. Lower R_{16} to 10 kΩ, increase C_{24} to 10 μF, and increase R_{15} to 270 kΩ.

3. Excessive heating of the transformer is caused by a d.c. component in the primary winding. This is simple to remedy by fitting a capacitor of any value between 47 nF and 470 nF, and a voltage rating of 630 V, across the primary connections. This capacitor is conveniently mounted on to the PCB terminal block that connects the transformer to the mains.

4. One final point: when using LED DVMs for the voltage/current indication, their ground line must be connected to the **positive** terminal of C_{12}.





Hard disk monitor

December 1989

In some cases, the circuit will not reset properly because the CLEAR input of IC_{3A} is erroneously connected to ground. Cut the ground track to pin 3 of IC_3, and use a short wire to connect pin 3 to pin 16 (+5 V).

Microprocessor-controlled telephone exchange

October 1990

In some cases, the timing of the signals applied to IC_{17} causes a latch-up in the circuit, so that the exchange does not detect the state of the connected telephones properly. Solve this problem by cutting the track to pin 1 of IC_{17}, and connecting pin 1 to ground (a suitable point is the lower terminal of C_6).

The text on the fitting of wires on the BASIC computer board (page 19, towards the bottom of the right-hand column) should be modified to read: 'Finally, connect pin 6 of K_2 to pin 7 of IC_3 ($\overline{Y7}$ signal).'

S-VHS/CVBS-to-RGB converter (2)

October 1990

The capacitor marked 'C37', next to R_{21} on the component overlay (Fig. 7b and ready-made printed circuit board), should be marked 'C39'.

In case they are difficult to obtain locally, inductors type 119-LN-A3753 (L_1) and 119-LN-A5783 (L_2) may be replaced with the respective types 119-ANA-5874HM and 119-ANA-5871HM, also from Toko, Inc. Suggested suppliers are Cirkit Distribution Ltd., and C-I Electronics.

EPROM simulator

December 1989

Counters IC_3 and IC_4 may not function properly owing to a too low supply voltage. This problem may be solved by replacing IC_{12} with a 7806. Alternatively, use BAT85 diodes in positions D_1 and D_2.

Programmer for the 8751

November 1990

The ready-programmed 8751 for this project is available at £35.25 (plus VAT) under order number **ESS 7061**, not under order number ESS 5951 as stated on the Readers Services pages in the November and December 1990 issues.

MEDIUM POWER A.F. AMPLIFIER

by T. Giffard

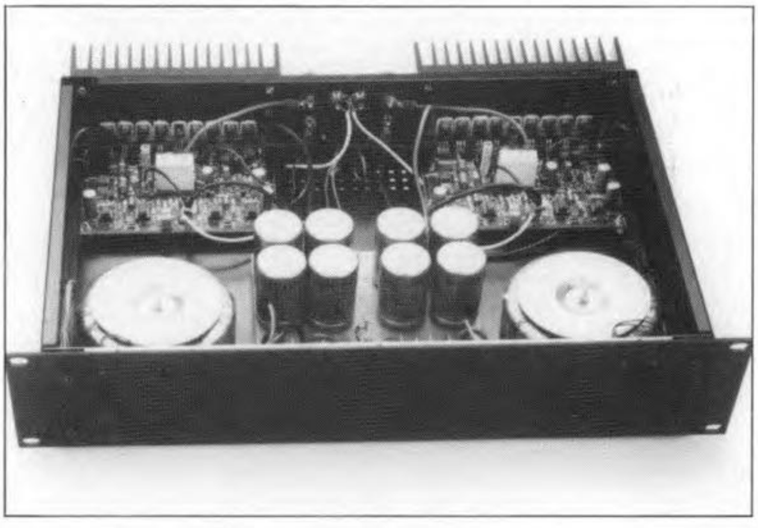
For the many readers who do not need hundreds of watts output from their audio hi-fi installation here is a modest 60-watt a.f. amplifier that is a match for even the best quality loudspeakers. Loads down to 2 ohms may be driven without any problem and with very low distortion. The design is entirely symmetrical and, apart from an input capacitor, direct coupled.

WHAT should be the power rating of a domestic hi-fi installation? Twenty watt, 50 watt, 200 watt? It is a vexed question that will never be answered to the satisfaction of every hi-fi buff. Psychology may play a role here, too: the difference in sound pressure between a 20-watt and a 200-watt system is only (!) 10 dB and that sounds a lot less than the difference of 180 watt in output powers. Be that as it may, a continuous power of 50–70 watt is more than adequate for at least 95% of all domestic hi-fi installations. It's far better to have a good-quality 50-watt system than a mediocre 200-watt one. Readers who take music reproduction seriously will, no doubt, have found this out themselves a long time ago. Having said that, it is admitted that there are a few, and fortunately only a few, loudspeakers that need at least a hundred watts to come to live.

Go to almost any hi-fi retailer and ask for a good-quality, medium power (say, 50 W) amplifier and you'll find that there are not many, if any. Until recently, the quality of almost all proprietary power amplifiers was in direct proportion to their power rating and that is the reason that many people buy a system with too high a power output for their requirements. Fortunately, some clever manufacturers have realized this anomaly and are doing something about it, so that amplifiers rated at 50–70 watt with a good specification are slowly becoming available. Nevertheless, we felt that a good-quality design for the enthusiastic DIY-er would be a welcome addition to the couple of high-quality preamplifiers we have published over the past few years.

Symmetrical design

Top priority in the design of the amplifier was quality, and this has resulted in a unit with a signal-to-noise ratio of not less than 100 dB at 1 W, harmonic distortion of not greater than 0.006% (8 Ω/50 W) and a slew rate of 30 V/μs.



TECHNICAL DATA
(Power supply 225 VA; buffer capacitor 20 000 μF per line)

Continuous power (THD = 0.1%)	60 W into 8 Ω		
	110 W into 4 Ω		
	170 W into 2 Ω		
Music power	200 W in to 2 Ω		
	(1 kHz; 20 ms on, 80 ms off)		
Harmonic distortion	50 W / 8 Ω	100 W / 4 Ω	150 W / 2 Ω
100 Hz	<0.0006%	<0.008%	<0.01%
1 kHz	<0.006%	<0.008%	<0.01%
10 kHz	<0.015%	<0.025%	<0.06%
Intermodulation distortion	<0.013% (1–50 W into 8 Ω at 1 kHz)		
	<0.05% (1 W into 4 Ω at 250 Hz–4 kHz)		
Slew rate	>30 V/μs (with input filter)		
Peak output current	15 A		
Signal-to-noise ratio	>100 dB at 1 W		
Output impedance	<0.02 Ω at 100 Hz and 1 kHz		
	<0.04 Ω at 10 kHz		
Input impedance	47 kΩ		
Input sensitivity	1 V r.m.s.		
Quiescent current	100 mA (nominal)		

If you imagine a horizontal line through the centre of the simplified circuit diagram in Fig. 1, the part above that line is a mirror image of that below the line (ignoring the DC correction and protection circuits).

The active DC correction circuit ensures that in no circumstances will direct voltage appear at the output, which is important in view of the direct coupling between all stages.

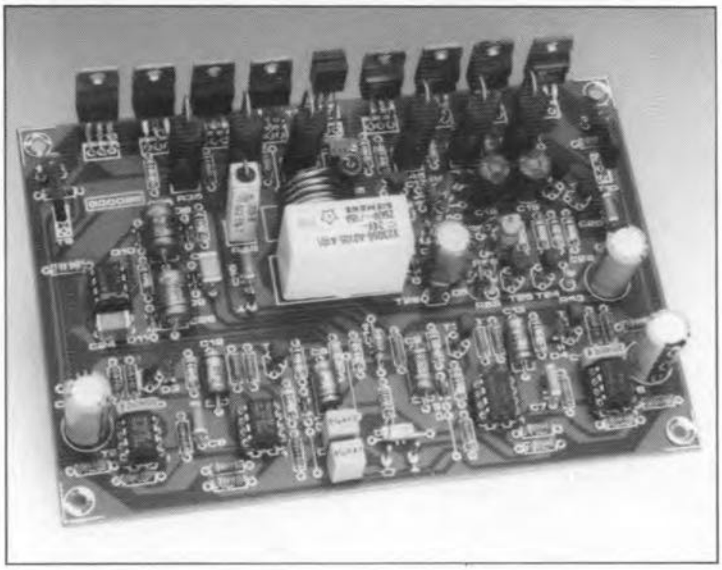
The protection circuit provides a delay on power-on, monitors whether there is any direct voltage at the output, and measures the current that is drawn by the output transistors. A mechanical relay is used, because we could not devise an electronic switching method that would satisfactorily limit the current without audible side-effects.

Although the design cannot be called revolutionary, we feel that parts of it are pretty original and combine some of the advantages of a number of other, standard designs.

The open-loop gain has been kept low, so that the amount of feedback can be kept small, which is all to the good of the overall quality. After all, the various stages then need contribute a smaller part of the overall amplification, which helps in keeping the distortion in each stage very small.

The input is formed by a differential amplifier, T1-T5, whose gain is limited to about 40 dB. The input transistors are coupled to another pair of differential amplifiers, T_2 and T_6 respectively, whose gain is about 27 dB. These are followed by controlled current sources T_9 and T_{10}. Setting of the quiescent current for the output stages is provided by a variable zener diode consisting of three transistors, which are connected between the collectors of T_9 and T_{10}.

The output stages consist of drivers T_{13} and T_{17}, and two pairs of three parallel-connected power transistors, each forming a super emitter follower.



Choice of components

In a symmetrical design, equality of the transistors in the input stages is of paramount importance. In our first design, self-paired BC transistors were tried, but these gave problems, particularly with thermal stability. It was therefore decided to use proprietary dual transistors, although these are rather more expensive. However, taking into account the necessary reliability of reproduction of the design, and looking at the near-perfect parallel operation of the dual transistors and their excellent thermal behaviour, the expense is well worth it.

Some of the problems a designer often has to cope with are concerned with the availability of components. The output transistors in the present design are a typical example. Although they were listed as standard types in a Philips catalogue, the ones we ordered had not been delivered six months later in spite of numerous telephone calls. Perhaps this explains why Philips has not been doing too well of late. In the end, the order was cancelled and replaced by one for SGS-Thomson devices (delivered in a few weeks). These are in most respects as good as the Philips types but lack somewhat in bandwidth, although in the prototypes that was not noticeable. On the other hand, the characteristics of the complementary n-p-n and p-n-p types are for all practical purposes identical.

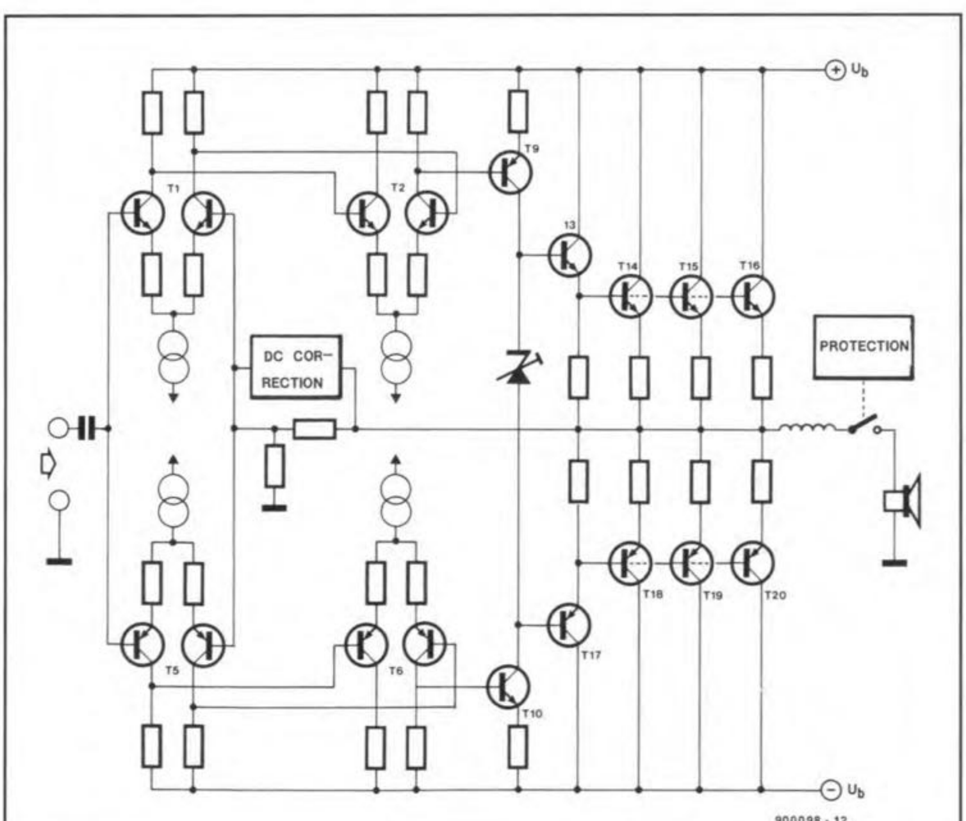


Fig. 1. The simplified circuit diagram of the amplifier clearly shows the symmetrical design.

The final design

The input of the amplifier—see Fig. 2—is protected against direct voltages by capacitors C_1 and C_2, which may be omitted if the preamplifier to be used already has a capacitor at its output. Be careful in future when connecting a different preamplifier which may not have a capacitor at its output.

The input capacitors are followed by a low-pass filter, R_1-C_3. This network limits the bandwidth of the input signal to obviate any transient intermodulation distortion (TIM). The cut-off frequency of the filter is about 180 kHz, assuming that the output impedance of the preamplifier is 50 Ω.

Dual transistors T_1 and T_5 form the symmetrical input stage, which is controlled by current sources T_3 and T_7. The current through each of the dual transistors is set at just above 1 mA. The gain of the differential input amplifiers is determined by the ratio of the values of their collector and emitter resistors. Networks R_5-C_6 and R_{14}-C_7 set the open-loop bandwidth at 500 Hz.

The second stage consists of dual transistors T_2 and T_6, which are controlled by current sources T_4 and T_8. This stage not only provides 27 dB gain, but als impedance matching between the input stage and the following current amplifier.

Current sources T_4 and T_8 use LEDs to obtain a voltage reference. Bear in mind that these diodes have a potential drop of about 1.6 V across them.

The quiescent current through current amplifiers T_9 and T_{10} is some 20 mA. This level of current is necessary to ensure that drivers T_{13} and T_{17} provide sufficient current in all circumstances. The amplification of these amplifiers is determined by the value of their emitter resistors and the input impedance of T_{13} and T_{17}.

The collectors of the current amplifiers are intercoupled via a presettable 'zener' network, which serves to set the quiescent operating point of the output stages. This network normally consists of just one transistor, but it was found that the thermal behaviour of this was not sufficient to compensate the thermal conduct of the output stages correctly and quickly enough. The three transistors finally chosen work almost perfectly and have the further advantage that they form a virtually ideal current source which ensures that current variations through them hardly affect the zener voltage.

The drivers and associated power transistors are, of course, of the same type. Three

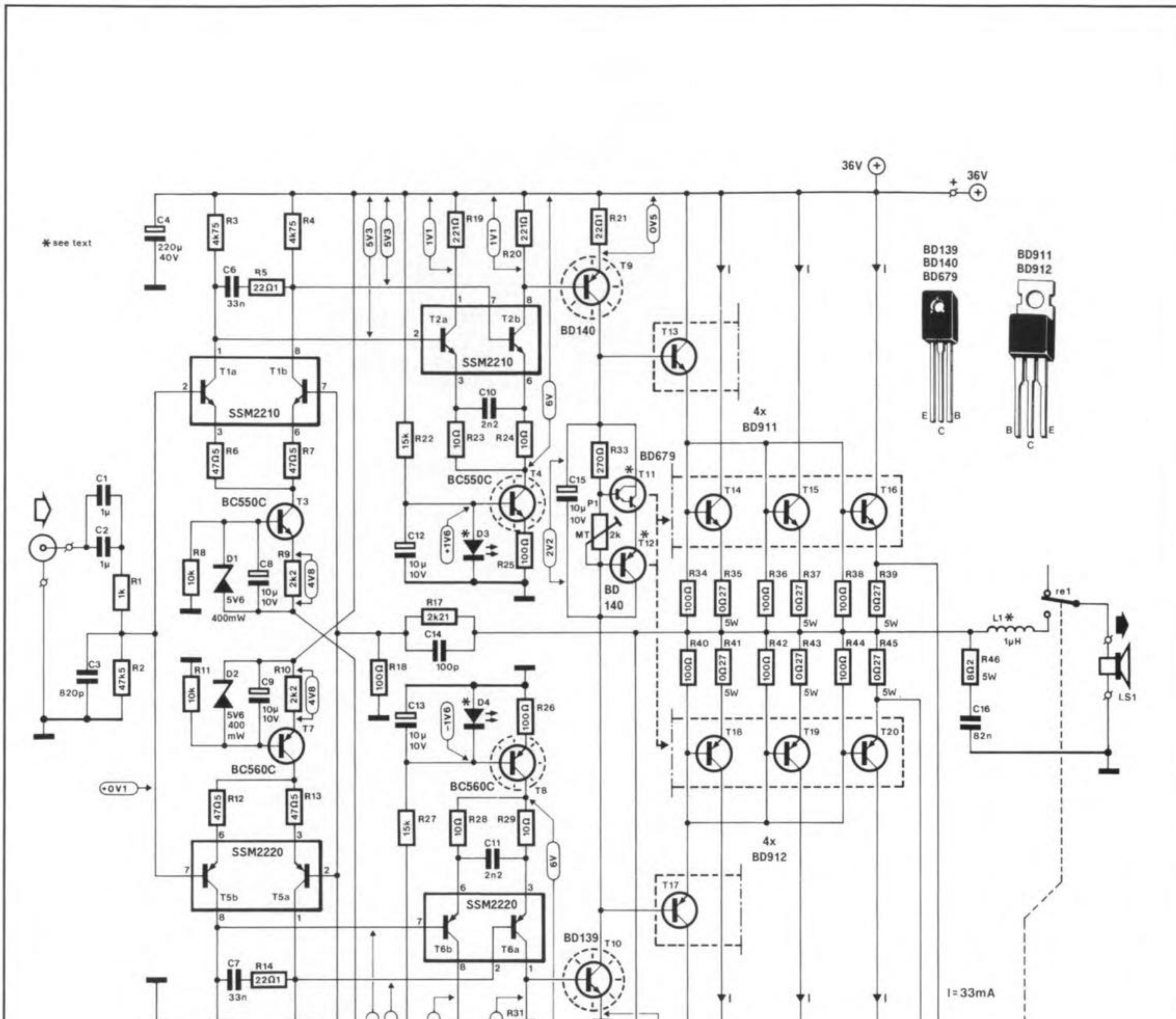


Fig. 2. Circuit diagram of the medium-power A.F. amplifier.

output transistors instead of one make for more efficient cooling; moreover, low-power transistors normally have better electrical properties than high-power ones.

A Boucherot-type network, R_{46}-C_{16}, at the output ensures that the amplifier remains adequately loaded at high frequencies. Inductor L_1 limits the rise time of the output signal when the load is capacitive. This results in slightly less damping at high frequencies, but even so the damping factor at 10 kHz and an output impedance of 8 Ω remains greater than 100.

Feedback is provided via resistors R_{17} and R_{18}, and an important role in this is played by DC correction circuit IC_1. Since the feedback loop does not include coupling capacitors, direct voltages at the input are amplified to the same degree as AC signals. Owing to the slightly asymmetric setting of the direct voltages at the differential input stages, which is unavoidable because of differences between n-p-n and p-n-p transistors and also in base resistance, it is essential that the output voltage of the amplifier is kept at 0 V, and this is effected by IC_1.

The output of the amplifier is applied to integrator IC_1 via low-pass filter R_{60}-C_{23}. The supply for the integrator is derived from the main 36-V supply with the aid of resistors R_{63} and R_{64}, and zener diodes D_{10} and D_{11}.

The protection circuit is formed by transistors T_{21}–T_{26}. When the supply is switched on, T_{26} is off, so that bistable T_{24}-T_{25} is reset, which results in T_{25} being switched on. Transistor T_{26} cannot conduct until its base-emitter voltage is high enough and this does not happen until C_{21} has been charged via R_{58}. It is this action that delays the actuation of the amplifier.

Once T_{26} is switched on, relay Re_1 is energized and diode D_7 lights to indicate that all is well.

Transistor T_{21} monitors the output current of the amplifier by measuring the potential drop across the emitter resistors of T_{16} and T_{20} via voltage divider R_{47}-R_{48}. If that current exceeds 5 A (that is, a total current throught the three output transistors of 15 A), T_{21} switches on which results in bistable T_{24}-T_{25} toggling after which the relay is deenergized within 5 ms.

Any direct voltage at the output is measured via low-pass filter R_{49}-R_{50}-C_{18}-C_{19}. If there is a direct voltage at a level of more than 1 V, T_{22} will switch on of the voltage is negative and T_{23} if it is positive. Again, the relay is deenergized via the bistable.

When the relay is denergized owing to too high a current or voltage, it will remain so until the supply is switched off. When, after a few seconds, the supply is switched on again, the relay will be energized if the fault condition has been removed.

The nominal supply voltage is 2×30 V, which may rise under no-load conditions to 2×37 V. Each amplifier may have its own power supply, but it is, of course, possible to power a stereo system (two amplifiers) from one supply only as discussed below.

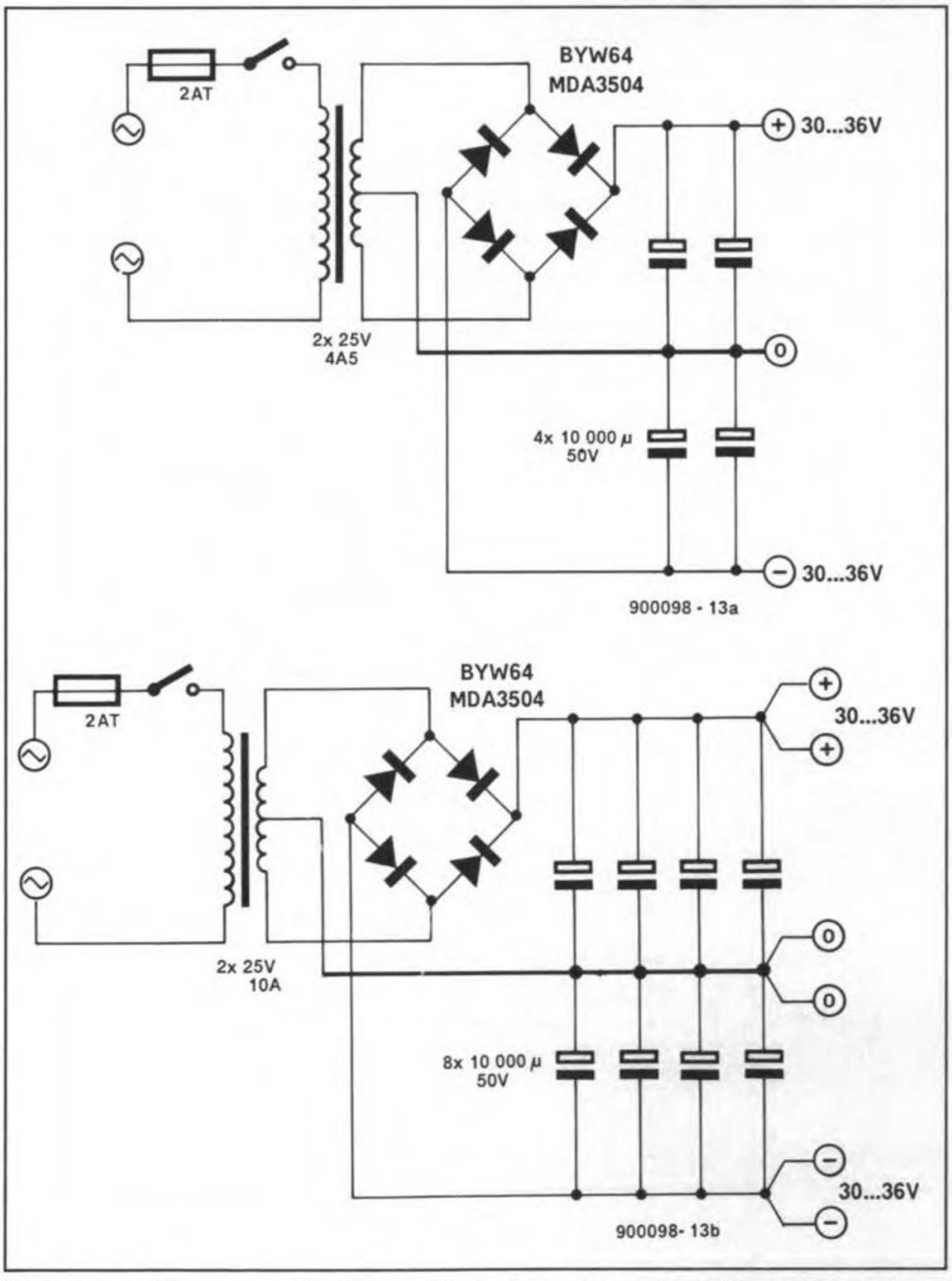


Fig. 3. Two possible power supplies: the one at the top is for a monaural amplifier, while the one below is a single supply intended for a stereo amplifier.

Power supply

In principle, there are three ways of providing the amplifier with power: a single supply for a monaural amplifier; a stereo amplifier with an independent supply for each amplifier; and a stereo amplifier with a single supply. The first two obviously provide the best possible channel separation.

The circuit of a power supply for a monaural amplifier is shown in Fig. 3a. The mains transformer specified provides sufficient power to allow the amplifier delivering continuous power into a 4 Ω load. It was not thought necessary to specify it for continuous power into 2 Ω. After all, the nominal resistance of many loudspeakers is 4 Ω, although there may be dips to 2–3 Ω. However, the electrolytic capacitors have sufficient capacitance to ensure adequate current during peaks in music reproduction. This explains why a total capacitance of 40 000 µF is specified.

The single supply for a stereo amplifier —see Fig. 3b—has a higher rated mains transformer. For normal use, a 6-A type will suffice. However, if low-impedance loudspeakers are used, it is better to use a 10-A type.

The electrolytic buffer capacitors are 50-V types; if these cannot be obtained, 63-V types may be used, although these are somewhat larger.

The construction of the amplifier will be described in next month's issue. ■

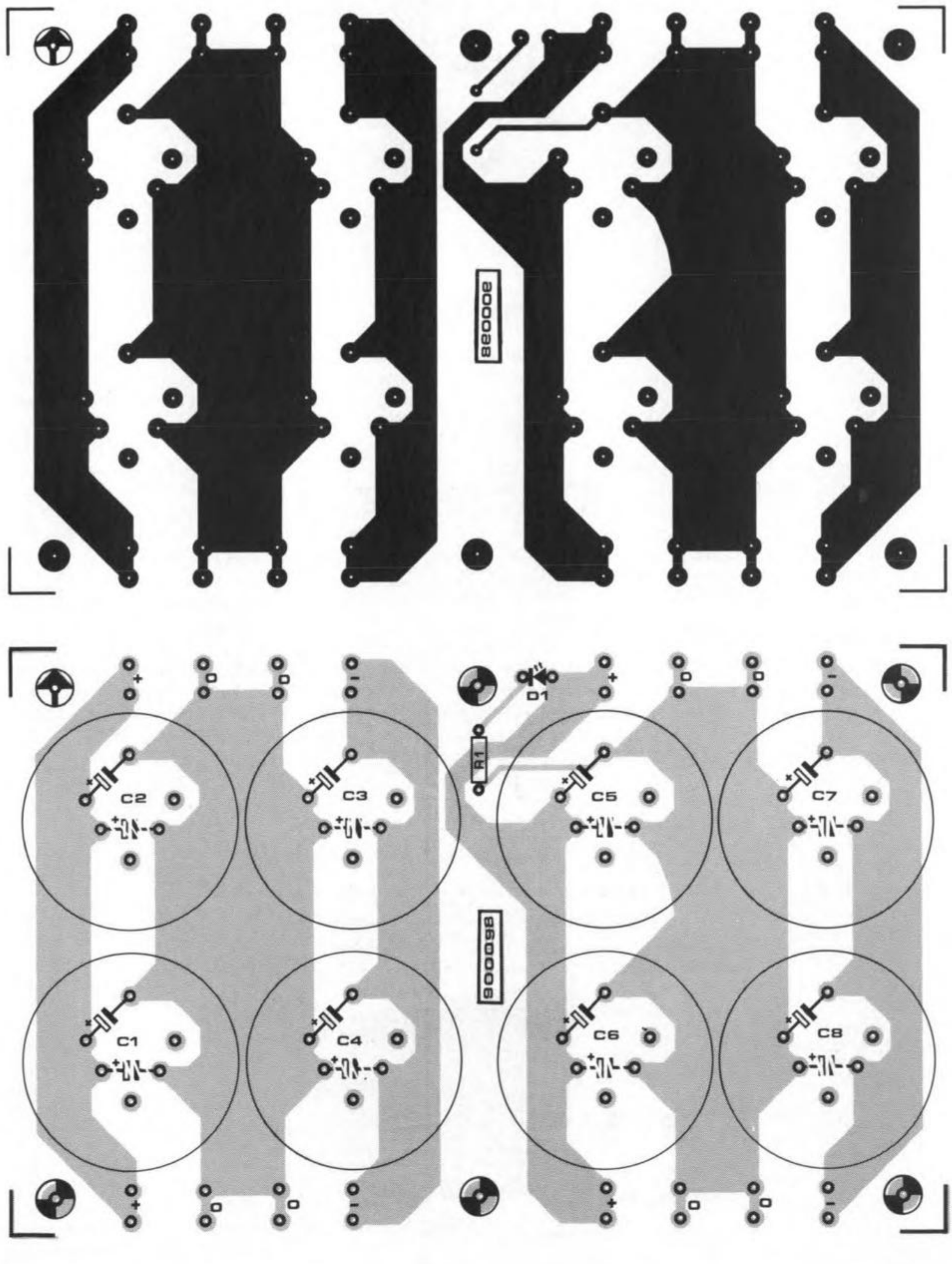
CTIONS CORRECTIONS CORRE

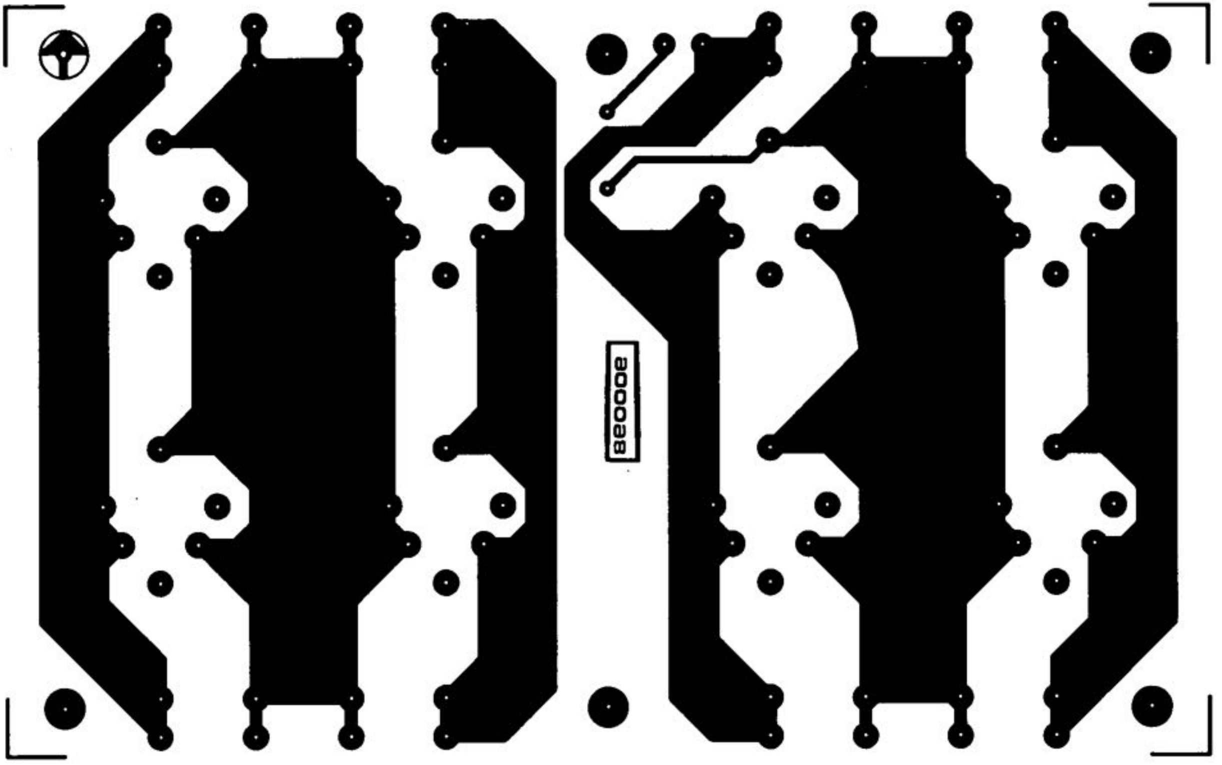
Medium-power A.F. amplifier

October and November 1990

A circuit board to accommodate the smoothing capacitors in the symmetrical power supply section of the amplifier was designed after publication of Part 2 of this project. The track layout (mirror image) and the component mounting plan are given in Fig. 3 to assist readers in producing this PCB, which is not available ready-made. ■

3





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DUBBING MIXER EV7000

PART 1: CONNECTION AND CIRCUIT DESCRIPTION

This sound mixer, designed and marketed as a kit by ELV, allows a variety of fading, sound dubbing and voice-over effects to be realized. The voice channel can override the music channel either automatically (by voice control) or manually. A total of fourteen controls and two toggle switches bring out all features of this easy-to-operate unit.

A dubbing mixer like the EV7000 is often used at parties and film or slide presentations, when a voice channel occasionally overrides the (background) music to provide announcements or comment. A smooth transition between the music and the voice channel requires a fader such as the one described here. To prevent different sound qualities on the music and the voice-over, the EV7000 has separate tone controls for each channel. In addition, separate volume and balance controls are provided.

Operation and controls

As shown by the above photograph, all indicators and control elements of the dubbing mixer are arranged on the front panel. The input and output connectors are located on the rear panel.

Before it is taken into use, the dubbing mixer is connected to the power supply and the available audio equipment. A small mains adapter with an output of 12 V at about 300 mA is connected to the 3.5-mm adapter socket on the rear panel. The 'Mic On' and 'Line on' LEDs on the front panel light to indicate that the mixer is on.

The dubbing mixer is best connected between the preamplifier and the power amplifier. The stereo output signal supplied by the preamplifier is connected to the phono (RCA-type) input sockets on the rear panel of the dubbing mixer. The outputs of the mixer are connected to the inputs of the power amplifier. When a mono preamplifier is used, its output signal is fed to the left input channel of the dubbing mixer, while the right input channel is not used.

The dubbing mixer has an internal ampli-

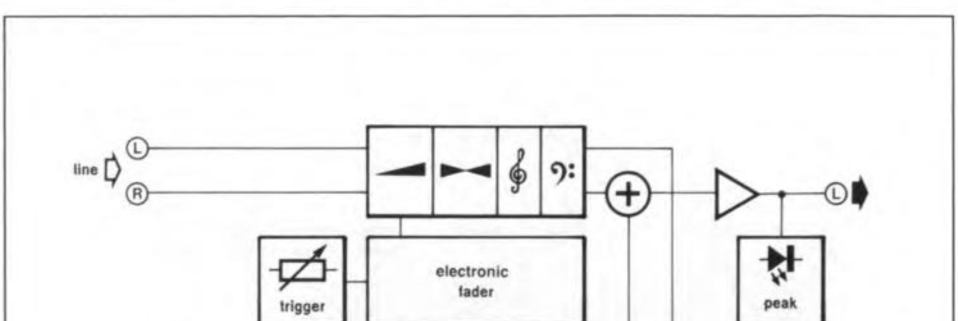

Fig. 1. Block diagram of the dubbing mixer.

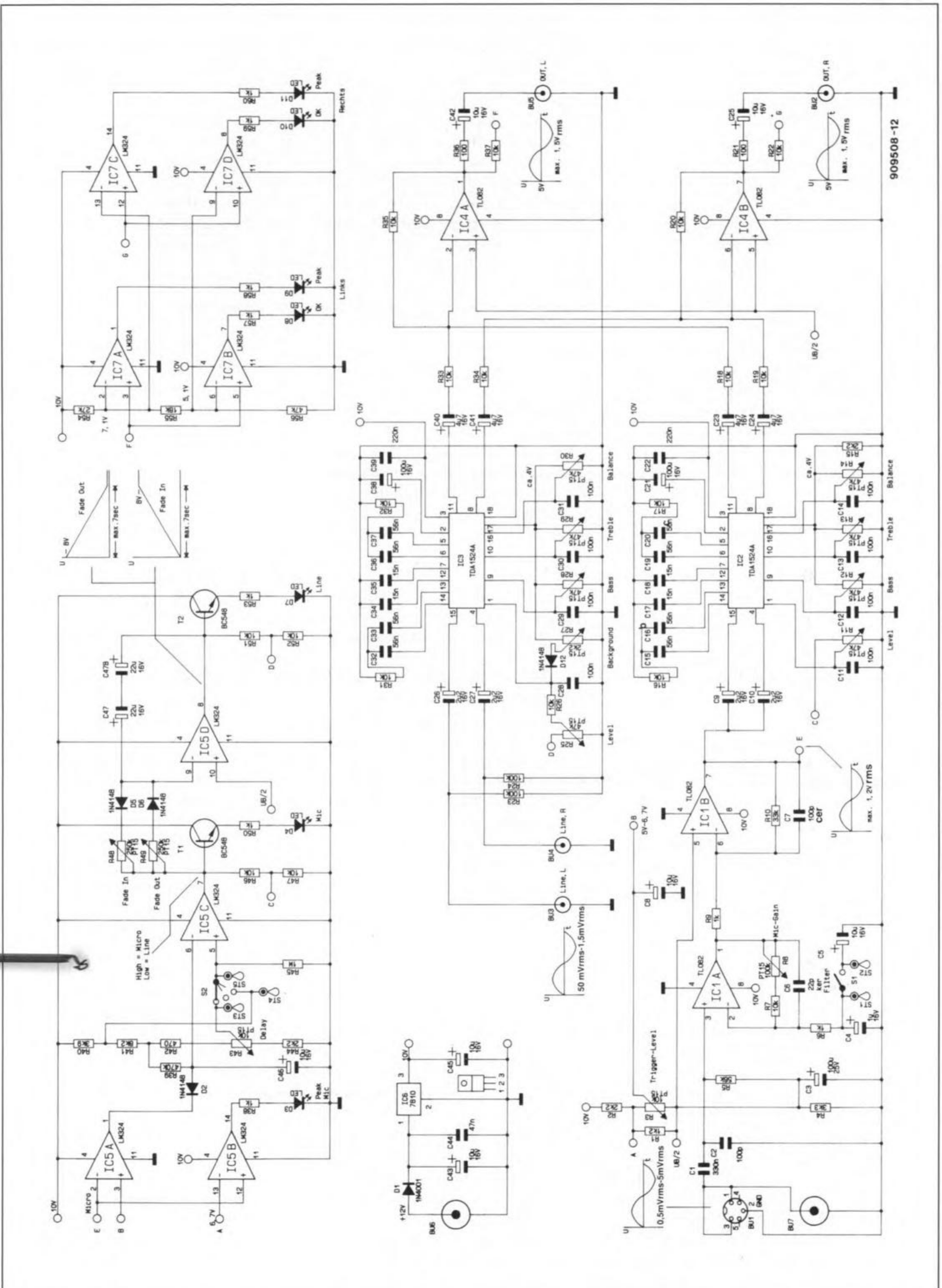

Fig. 2. Circuit diagram of the dubbing mixer. The central parts are two audio signal processing ICs Type TDA1524A (IC2; IC3).

fier capable of handling input levels between about 50 mV_{rms} and 1.7 V_{rms}, so that virtually any (line-) signal source can be connected.

The stereo output signal of the dubbing mixer has a maximum level of about 1.7 V_{rms}, and can be set to the required volume within a range of 100 dB. This allows the mixer to drive almost any type of power amplifier.

The mixer has two parallel-connected microphone input sockets: one for a 3.5-mm jack plug and one for a DIN plug.

The controls on the front panel are arranged into three areas, marked by white lines. Much of what follows below on the basic operation of the mixer is illustrated by the block diagram in Fig. 1.

- The controls for the **microphone** channel are located in the top left-hand area on the front panel. The LED at the extreme left indicates the peak microphone level. The tone and volume controls are only effective for the microphone channel. The toggle switch marked 'function' allows three different modes to be selected:
  – 'off' to disable the microphone channel;
  – 'on' to enable the microphone channel;
  – 'auto' to enable the automatic voice control (VOX) function. When selected, this function provides an automatic fade-in when a certain (preset) microphone level is exceeded. The microphone is faded out automatically when its signal level drops below the preset level. The state of the VOX is indicated by the 'Mic on' LED.
- The controls for the **line** channel are found in the lower left-hand area on the front panel. The four LEDs at the left are used as level indicators for the two output channels. The 'Pe' (peak) LEDs should preferably remain off, while the 'OK' LEDs light when the signals are at a sufficiently high level. The minimum recommended signal level is indicated by the 'OK' LEDs flashing irregularly. Like the microphone channel, the line channel has separate tone controls and a level control, marked 'background'. The latter sets the background level of the music signal while the voice channel is actuated. When set fully counter-clockwise, the music is totally suppressed during the voice-over.
  The 'Line on' LED indicates the fade-in and fade-out actions. The intensity at which the LED lights is a rough indication of the background music level during the voice-over. When the LED lights at its full intensity, the music channel is at maximum level, i.e., the microphone channel is off.
- The **basic functions** are set by the controls in the right-hand area on the front panel. The 'Mic gain' potentiometer determines the amplification in the microphone channel. It should be set to a position at which the microphone signal is loud enough when the 'Mic level' control is in the last one-third of its travel. The toggle switch marked 'Mic filter' allows the low side of the frequency range to be limited to about 200 Hz. This mode is particularly suited to the suppression of floor noises, rumble and other low-frequency interference. When the filter is switched off, the frequency range starts at about 20 Hz, which makes the microphone channel usable even for music signals.
  The 'trigger level' control sets the threshold of the previously mentioned VOX function. Turning this control clockwise results in a higher switching threshold, i.e., a higher microphone signal level at which the dubbing mixer switches to voice-over. The 'trigger level' control is, however, enabled only when the 'function' switch is set to the 'auto' position.
  The three timing controls in the lower right-hand corner allow the speed of the fade-in and fade-out effects to be changed to requirement. The 'fade-in' control has a range of 0 to about 7 seconds, independent of the time set with the 'fade-out' control. The 'delay' control determines the time between the end of the voice-over and the start of the music channel fade-in. The range of this control is 0 to about 5 seconds. Like the VOX level control, the 'delay' control is active in the 'auto' mode only. It is intended mainly to prevent the music channel being faded in during every short pause in the voice channel.

Circuit description

The crucial parts in the circuit diagram in Fig. 2 are two audio processor ICs Type TDA1524A. All signal parameters (volume, balance, and tone) are set by control voltages.

The left and right line signals are applied to sockets BU3 and BU4 respectively and fed to input pins 4 and 15 of the TDA1524A (IC3) via coupling capacitors C26 and C27. Since all active parts are contained in the TDA1524A, only a handful of external capacitors and resistors is required to achieve the signal conditioning functions. The bass level is determined by R31, C32 and C33 for the left channel, and R32, C36 and C37 for the right channel. The treble controls require one capacitor only: C34 (left) and C35 (right). The bass and treble settings are controlled by electronic potentiometers in the TDA1524A. These potentiometers, in turn, are controlled by externally applied direct voltages. The volume is set via control input pin 1, the bass level via pin 9, the treble level via pin 10, and the balance via pin 16. At a supply voltage of 10 V, the range of the control voltage is about 0.25 V to 4.0 on all these inputs. The level of the supply voltage hardly affects the settings, however, since the potentiometers that set the sound parameters, R27 – R30, are connected to the reference voltage supplied by pin 17 of the TDA1524A. Capacitors C28 – C31 serve to suppress contact noises as the potentiometers are turned.

The two volume controls that operate at pin 1 of IC3 form a special configuration, in which diode D12 is an important component. The control voltage at the wiper of potentiometer R25 is fed to pin 1 of IC3 via R26 and C28. The positive supply voltage for the potentiometer is provided by the output of IC5 and potential divider R51-R52. In the 'line' mode, R25 is supplied with about 4 V, which allows the full volume range to be covered. When the microphone is switched on, the potentiometer supply voltage drops to about 0 V, which would normally result in total suppression of the music signal. Diode D12, however, keeps the minimum volume control voltage at 0.7 V below the level set with R27, the 'background' level control. The operation of the driver circuit around IC5 will be reverted to in due course.

Electrolytic capacitor C38 forms a buffer for the internal supply voltage of the TDA1524A, while C39 filters the supply voltage applied to the circuit.

The output signals at pin 11 (left) and 8 (right) of the TDA1524A are fed to the summing inputs of inverting amplifiers IC4A and IC4B via C40-R33 and C41-R34. The summing operation involves the two line signals (left and right) and the microphone signal supplied by IC2. Components with values equal to those used for the line signals take the microphone signal from output pins 8 and 11 of IC2 to input pins 2 and 6 of IC4. The two opamps, IC4A and IC4B, serve to invert and to buffer the signals.

The left channel signal reaches the output socket, BU5, via the output of IC4A, pin 1, and R36-C42. Similarly, the right channel signal arrives at BU2 via pin 7 of IC4 and R21-C25.

The microphone signals are treated by IC2 and potentiometers R11 to R14, in a manner similar to that described for the line amplifier, IC3. The supply voltage of the volume control potentiometers is provided by pin 7 of IC5C and potential divider R46-R47. The two inputs of IC2 are driven in parallel by the output of opamp IC1B. ❑

(to be continued next month)